Electronic technology

principal contents

Nyquist Diagram Tracer for A.F.

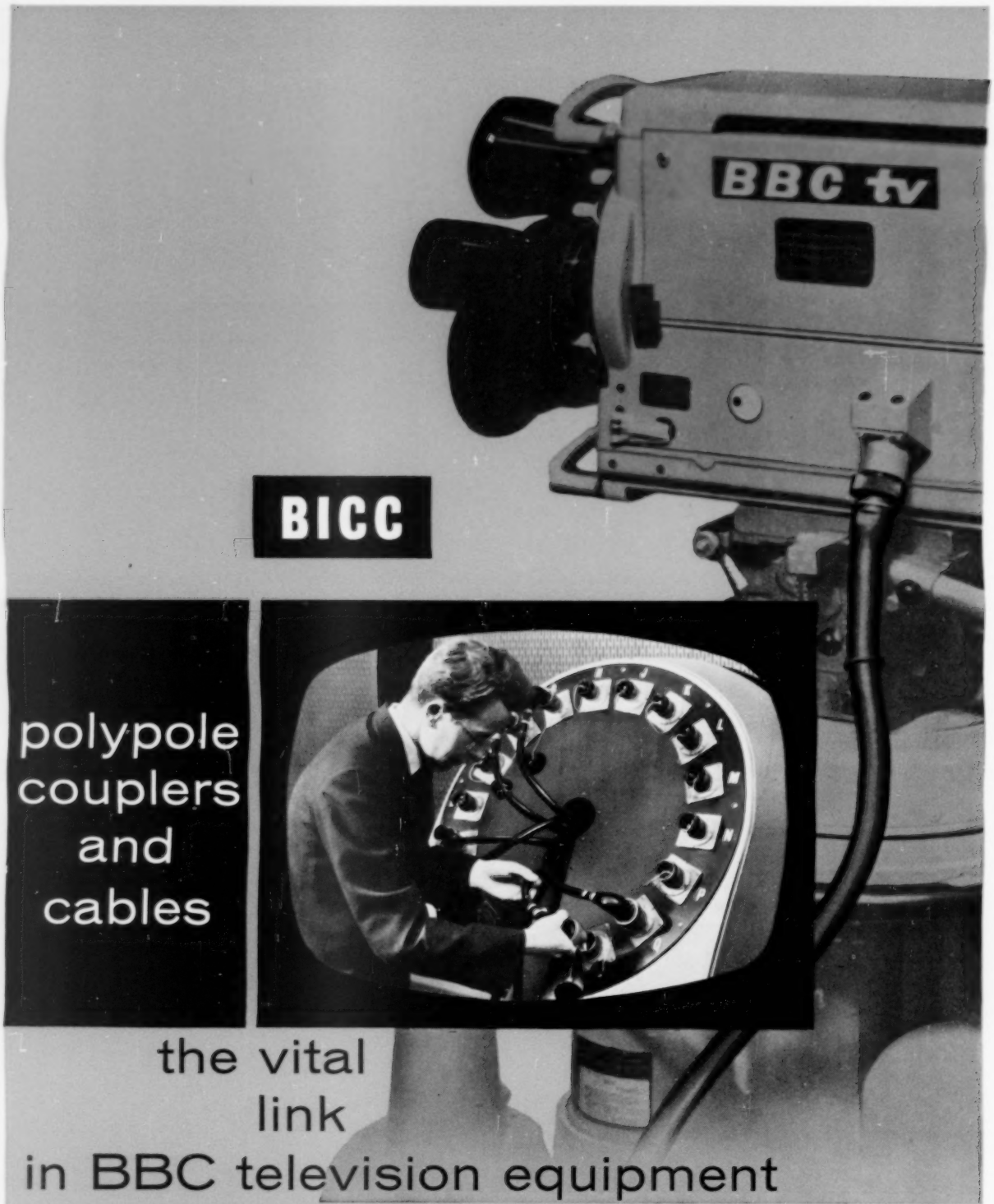
Two-Element Aerial Array

Bandpass Transistor Amplifiers

The Round-Travis Discriminator

FIVE SHILLINGS

MAY 1961 VOLUME 38 NUMBER 5



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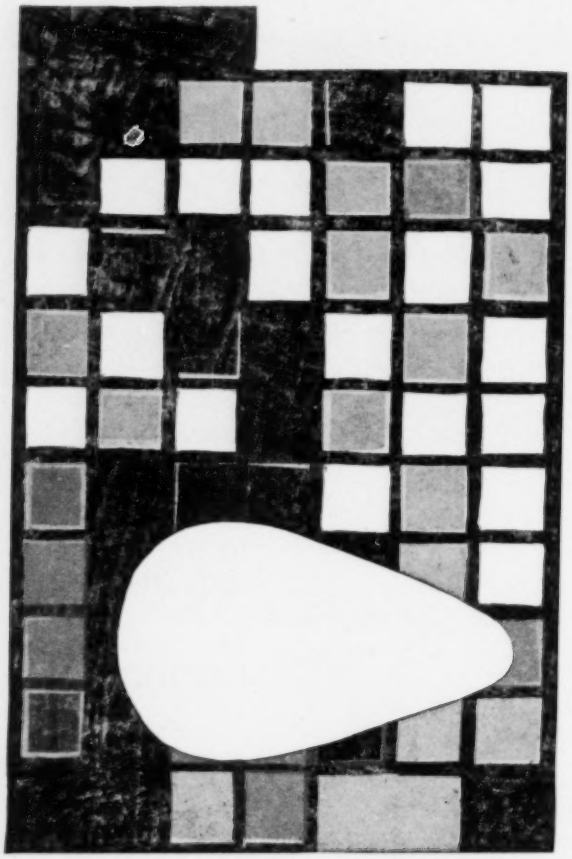

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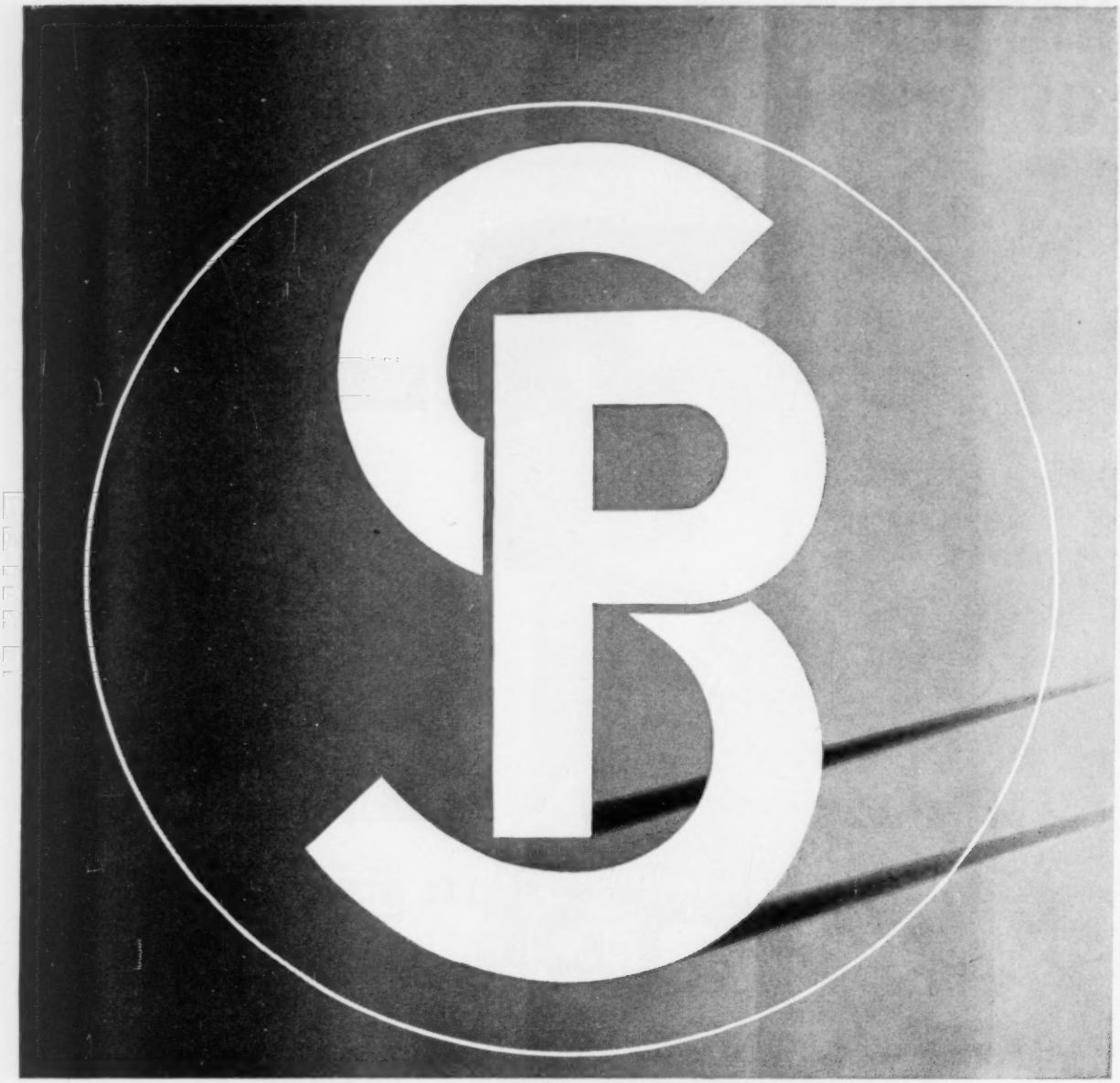

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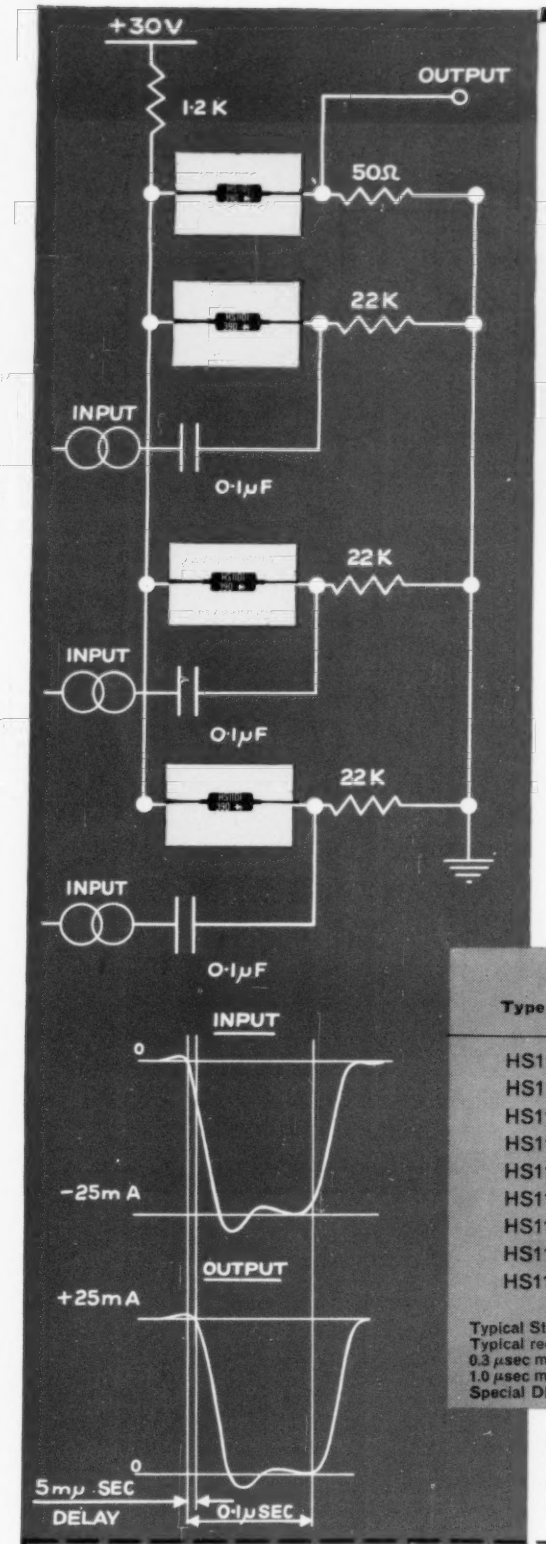

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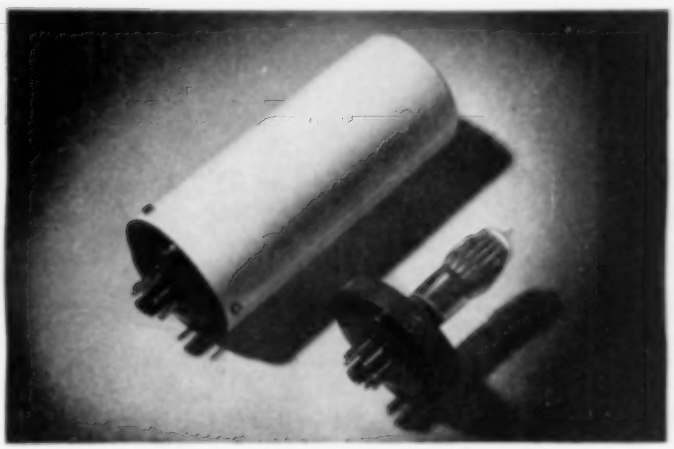

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CONTENTS

VOLUME 38 NUMBER 5 MAY 1961

Circuit Diagrams	**155**	Editorial
Nyquist Diagram Tracer for A.F.	**156**	*by Arthur R. Bailey, M.Sc.(Eng.)*
Two-Element Aerial Array	**160**	*by V. G. Welsby, Ph.D.*
New Distortion Criterion—Part 2	**163**	*by E. R. Wigan, B.Sc.(Eng.)*
Mathematical Tools	**175**	*by Computer*
Theory of Waveguides and Cavities—Part 3	**178**	*by R. A. Waldron, M.A.(Cantab.)*
Bandpass Transistor Amplifiers	**183**	*by C. J. McCluskey, B.Sc.*
The Round-Travis Discriminator	**186**	*by K. R. Sturley, Ph.D.*
Correspondence	**188**	
New Books	**189**	
Standard-Frequency Transmissions	**191**	
New Products	**192**	
Abstracts and References	**A67–A82**	

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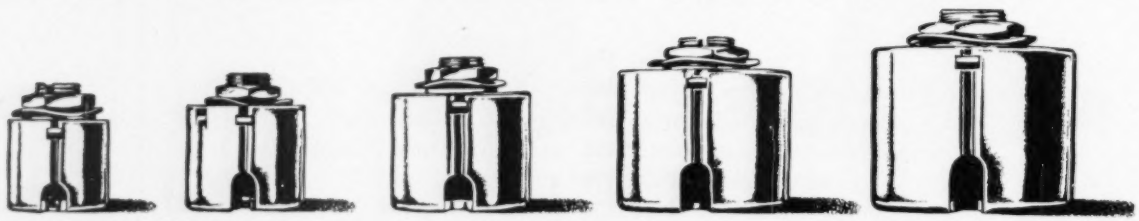

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Circuit Diagrams

IN the drawing of circuit diagrams of valve apparatus a more-or-less standard practice has grown up over the years. This is to draw valve symbols with anode at the top and supply-voltage busbars increasingly positive upwards. As a result, many circuits are recognizable at a glance without study; that is needed only when the relatively minor details must be understood.

As a correspondent points out elsewhere in this issue, a difficulty arises with p-n-p transistor circuits. If the standard of line-voltage busbars is adhered to, the rest of the circuit is upside down compared with valve practice. If the collector is treated as analogous to the anode of the valve and drawn at the top, the main part of the circuit comes right way up but the supply lines become increasingly negative upwards.

We ourselves regard the direction of polarity of the supply lines as a triviality compared with ease of recognition of the bulk of the circuit. Our own opinion is, therefore, that transistor circuits should be drawn valve way up, with supply lines increasingly negative upwards.

This is also our correspondent's view and, in his letter, he indicates a solution to some of the minor problems that arise, as when it is desired to include waveform diagrams.

From quite early in the history of the transistor, there have been two schools of thought about this. As time goes on, each becomes more accustomed to its own preference and more intolerant of the other. Sooner or later standardization must come and the sooner the better, since fewer people will have to change their habits.

Up to the present we have not imposed any standard on our contributors but have allowed them to adopt their own preferences in articles in *Electronic Technology*. As a result, circuits drawn in both ways have appeared in our pages. We have done this so that both ways might be given a fair trial.

The proof of the pudding is in the eating. We find that to understand anything but the simplest transistor circuit, we have to re-draw it valve way up. Laborious as this may be, it is quicker than trying to understand it upside down. It usually then turns out to be something quite ordinary which, valve way up, we should have recognized for what it was at a glance.

No doubt, advocates of the 'positive-upwards' convention will say that with practice we should recognize circuits equally well in their convention. That is true and would carry great weight if the p-n-p transistor had displaced everything else. The valve is far from dead, however, and the n-p-n transistor does exist. If the 'positive-upwards' convention is retained it means that everyone will have to learn to recognize circuits at a glance, whichever way up they are drawn. This does seem an unnecessary burden to impose upon people.

Nyquist Diagram Tracer For A.F.

CIRCUIT AND PERFORMANCE DETAILS

By Arthur R. Bailey, M.Sc.(Eng.), B.Sc.(Eng.), A.M.I.E.E.*

SUMMARY. *The instrument described in this article enables the shape of the Nyquist diagram of a circuit to be determined in a very short time. The frequency range covered is from 20 c/s to 5 kc/s, readings up to 20 kc/s being obtainable if suitable magnitude corrections are applied. The output may also be readily displayed as the resolved components of the signal under test by the use of two d.c. meters.*

In work connected with complex a.c. networks, the determination of the transfer function of the circuit is usually required. This transfer function can be represented in many ways, the complex polynomial expression being often the most convenient mathematical expression. For most applications however, the transfer function form is more easily understood by using graphical methods.

The forms of presentation most commonly used are gain/frequency and phase/frequency plots (Bode diagrams), the Nicholls diagram and the Nyquist diagrams. Whichever method is used there exists the necessity of taking two readings for each increment in frequency. The process of evaluating the performance of a system is tedious, even if only for a single plot, but when it is desired to see the effect of varying circuit constants the problem can become very serious. It was for this reason that this particular instrument was developed, the time taken in development being far outweighed by the speed at which circuit performance can be determined.

The Nyquist diagram was chosen as all the scales are linear and where problems of stability are encountered the gain and phase margins are immediately obvious. Also the shape of the curve obtained will give a good idea of the characteristics of the system by comparison with the Nyquist diagrams of circuits already known.

The basic circuit is quite straightforward and is shown in Fig. 1. The reference test signal being fed into the circuit under examination is also fed into two squaring amplifiers, one of these incorporating an inherent 90° phase shift. The output from these squaring circuits is amplified and is then used to gate phase-sensitive rectifiers. The output from these rectifiers is d.c. amplified, the a.c. components filtered out, and the resulting d.c. is used to deflect the spot on the screen of a cathode-ray tube. The output of the circuit under test is fed to the inputs of the modulators, the d.c. output of the two channels, therefore, representing the resolved components (in-phase and quadrature) of the output of the circuit. The spot on the screen of the cathode-ray tube, therefore, deflects along its two axes for a distance proportional to the magnitude of the output of the circuit, one axis representing in-phase, the other quadrature voltage.

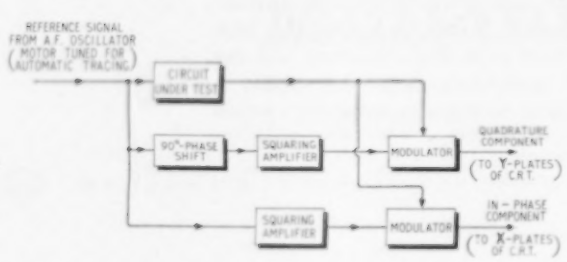


Fig. 1. *Basic circuit arrangement*

Square-wave gating was used on the modulators as this ensures that the output of the modulators will be independent of the gating input, providing that this is sufficient to turn the modulators from 'Full on' to 'Full off'. Linear-product modulators such as heptodes, used with sine-wave drive, would not have this advantage. The disadvantage of square-wave gating is that the circuit will give an output with odd harmonics fed to the input, even if no fundamental component is present. The output, however, will vary inversely with the harmonic number so this will normally cause little trouble. The effect of hum or similar noise that is non-synchronous with the supply is to give an a.c. output at the difference frequency between this frequency and that of the test signal. Providing these frequencies are more than a few cycles-per-second apart the filter circuits will remove them. At lower frequencies some of the output will pass on to the deflector plates and will make the spot circle its correct position at the difference frequency. This makes reading difficult, but this will only happen with very large interfering voltages and the correct value can be determined by estimating the centre of the circle.

A full block diagram of the arrangement is shown in Fig. 2, where it will be seen that push-pull operation is used for the modulators. This is necessary if the

* *Senior Lecturer in Electrical Engineering at the Bradford Institute of Technology.*

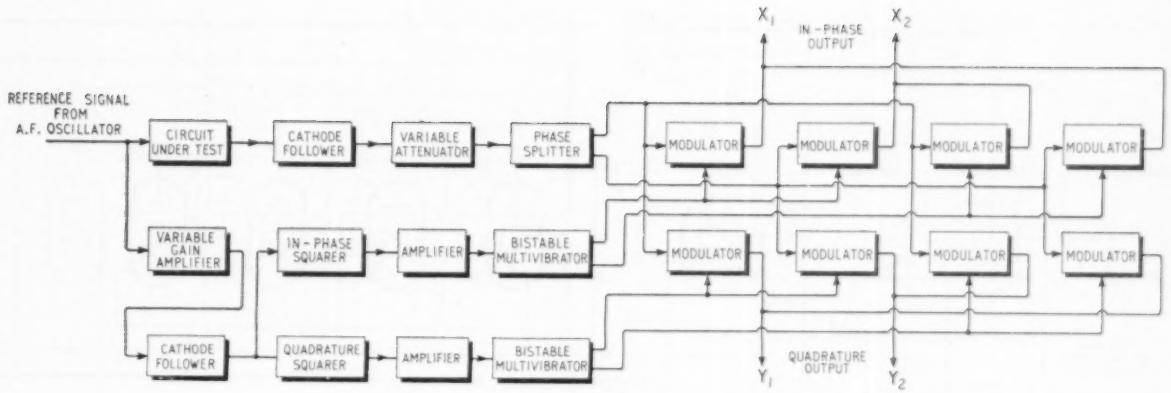


Fig. 2. *Complete block diagram, excluding power supplies*

circuit is to be drift-free, as single-ended d.c. amplifiers are very prone to drift. The problem of removing the a.c. components at the anodes is also eased at the same time. The d.c. amplifiers are, in fact, the same valves as the modulators; pentodes gated on their suppressor grids.

The circuit diagram of the complete equipment is shown in Fig. 3. The circuit component values are nowhere critical, the only precaution necessary being the provision of reasonably-balanced valves for the modulators. Excessive unbalance is indicated by the failure of the zero-setting potentiometers to centre the spot on the screen. Commencing at the reference input on the circuit diagram, the signal is first fed into a potentiometer, the setting of which is not critical but will allow for widely varying reference-signal levels. The output is then amplified by V_{11} and fed to the cathode-follower stage V_{12}. This valve is a power tetrode as the quadrature squaring diodes take a rather high a.c. current at the high frequencies. The output from V_{12} feeds the four squaring diodes D_5 D_6 D_7 and D_8. The diodes used are silicon junction diodes, these being used on account of their very rapid transition from the conducting to non-conducting states and their high ratio of conducting to non-conducting resistance.

The in-phase squaring diodes are fed via a resistance (220 kΩ) and the output square wave of approximately 1 V peak-to-peak, therefore, crosses the zero line at approximately the same instant as the input sine-wave drive. The output is approximately square by virtue of the output from V_{12} being in the region of 100 V peak-to-peak.

The quadrature squaring diodes operate in a similar manner except that their current is controlled by a capacitor and not a resistor. This means that if the input voltage is so high that the diode drop can be ignored the current in the capacitor and the diodes will lead the applied voltage by 90°. The diode output square wave will therefore cross the axis 90° out-of-phase with the applied sine-wave drive. In fact, errors of up to 1° have been measured, but this was considered satisfactory for the purpose in mind. This circuit has great advantages over the integrator/limiter circuit both from a cost viewpoint and also that the frequency range covered is very wide with little error. Circuit design is also much easier as hum problems are absent. The high-frequency range is apparently limited by junction capacitance, the low-frequency range by the 'insulation resistance' of the diode. Even so, the circuit will cover a frequency range of about 100–1 without excessive error. So far as the author is aware, this circuit has not been used before and it makes the design of this instrument much easier.

Two capacitors are used for the quadrature squaring diodes, thus giving the complete frequency coverage of the instrument in two widely-overlapping bands. The centre of the overlap frequency is at about 250 c/s.

The outputs from the squaring circuits are fed to two similar amplifiers V_{13} and V_{16}, the outputs being fed to the two bistable circuits comprising V_{14}, V_{15} and V_{17}, V_{18}. These circuits ensure that there is a very rapid changeover between opposite half-cycles of the square waves and also provide the required push-pull output for driving the modulators. These modulators V_3, V_4, V_5 V_6 and V_7, V_8, V_9, V_{10} are gated on their suppressor grids by means of the square-wave output from the bistable circuits. The germanium diodes D_1, D_2, D_3 and D_4 prevent the suppressor grids from being driven positive, in effect, clamping the excursion of the grids in the positive direction to approximately cathode potential. The high reverse resistance of these diodes also gives the suppressor-grid circuits the long time-constant required for low-frequency operation.

The output signal from the circuit under test is first fed into the cathode-follower stage V_1. This gives a very high value of input resistance (approximately 10 MΩ) and also provides a low output impedance suitable for driving the attenuator. The resistor and capacitor network associated with the attenuator is inserted to cancel spurious phase shifts above 5 kc/s, the values being determined experimentally. This gives a magnitude error as shown in Fig. 4, but it was felt that a single correction was better than two separate smaller ones.

The signal from the cathode-follower is then fed to a concertina phase-splitter V_2, which provides the required push-pull output for the control grids of the modulated pentodes.

The modulator valves V_3, V_4, V_5, V_6 and V_7,

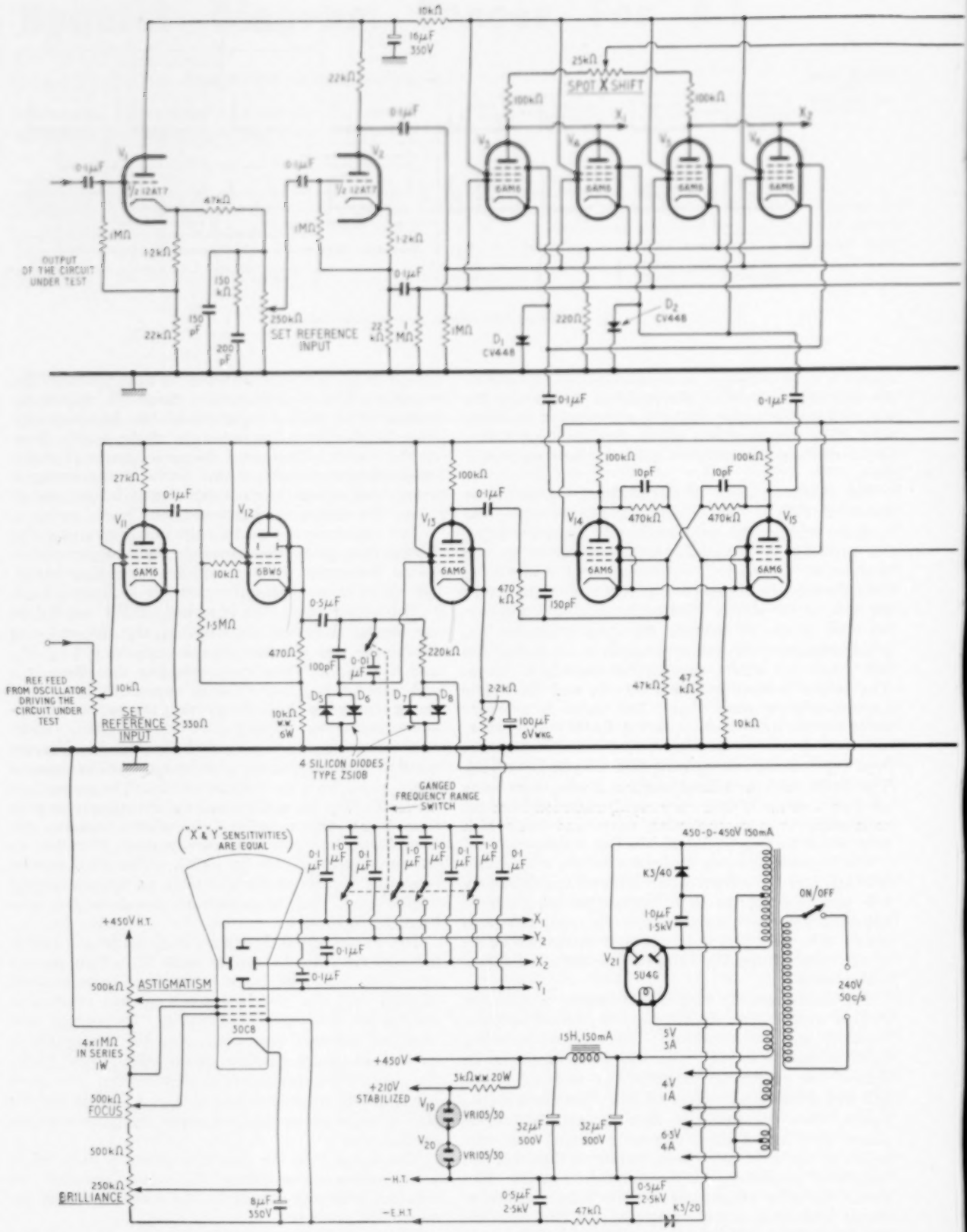
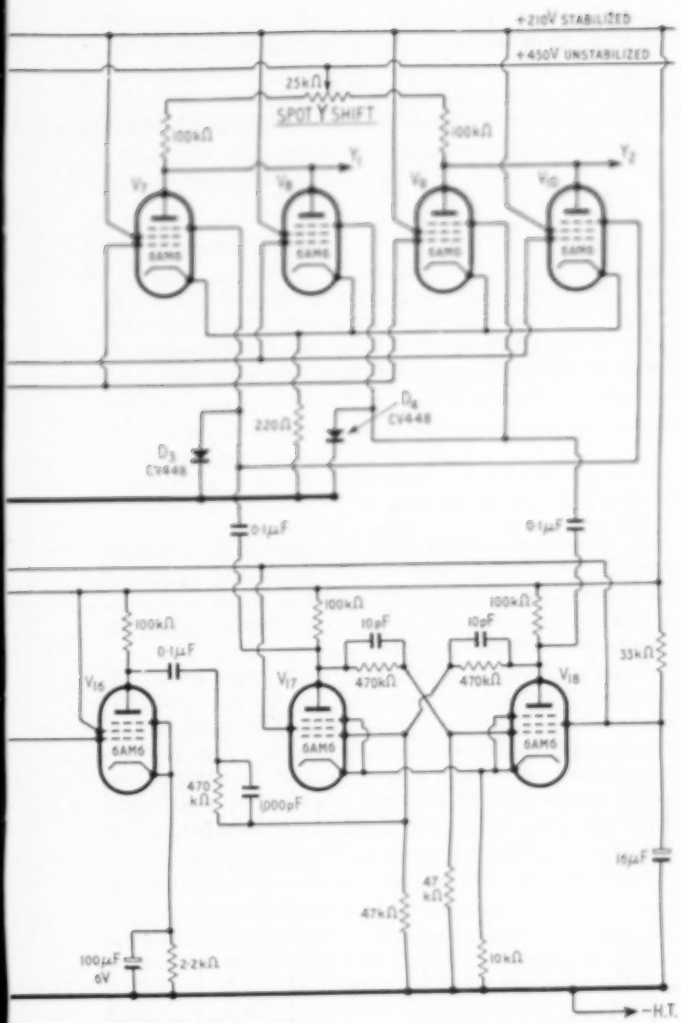


Fig. 3. *Complete circuit of the automatic Nyquist diagram tracer, excluding motor-tuned a.f. oscillator*



V_8, V_9, V_{10} are connected in balanced pairs and have unbypassed cathode resistors. This gives some reduction in the push-pull a.c. components as well as balanced drift in the output. The anodes of the modulators are decoupled to earth so as to reduce the a.c. components of anode voltage and so reduce any tendency to overloading due to excessive anode-voltage swings. Additional capacitors between the anodes of the valve pairs are included for the same reason. These capacitors also reduce the a.c. fed to the cathode-ray tube deflector plates to a very small level. This is necessary or the small spot will become blurred or even trace out a pattern due to the a.c. components. It is for this latter reason that additional capacitance is placed across the outputs from the modulators when switched for low-frequency operation. These capacitors do however give the instrument a time-constant of the order of several seconds, so they are removed as being undesirable at the higher frequencies.

The push-pull output from the modulators is fed directly to the deflector plates of a cathode-ray tube. These voltages give the spot a deflection along the 'X' axis proportional to the in-phase component from the circuit under test, the 'Y' deflection being proportional to the quadrature component. The deflection of the spot from its central zero position is therefore directly proportional to the magnitude and phase of the input signal. The screen can be calibrated in rectangular co-ordinates if an $x + jy$ form is required or, alternatively, in polar co-ordinates if the normal form $(r \angle\theta)$ of the Nyquist diagram is preferred.

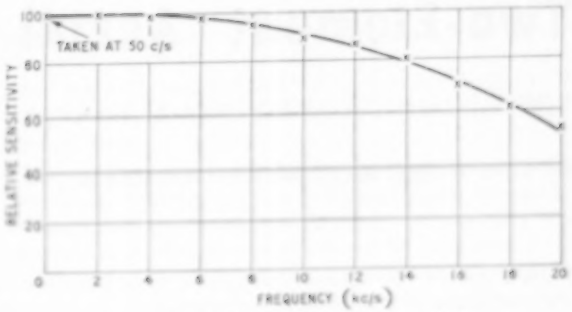


Fig. 4. Overall deflection sensitivity ('X' and 'Y').

If a meter indication of the resolved components is required rather than a display on a cathode-ray tube, then this is easily obtained by connecting centre-zero 100 μA meters between the pairs of modulator anodes. The cathode-ray tube and its associated supplies will be unnecessary in this case, thus giving a saving in both space and cost. The resulting resolved-components meter will not have quite such a wide frequency range as some commercially available, but it has the great advantage of not requiring resetting for varying frequencies of the test signal.

The performance details of the complete instrument are as follows:

Maximum sensitivity ..	10 cm/V per channel.
Frequency range.. ..	20 c/s to 5 kc/s, usable up to 20 kc/s with corrections.
Accuracy—magnitude ..	± 2%.
Accuracy—phase ..	± 2° up to 5 kc/s ± 4° up to 20 kc/s
Zero stability	Less than 1-mm shift for reference frequencies up to 5 kc/s Zero resetting is required above this frequency.

The instrument was originally designed for the investigation of non-minimum phase networks in connection with the research being carried out by the author and approved by the University of London for presentation for the degree of Ph.D. In fact, the instrument has been put to much wider use and enables a great variety of phenomena to be readily demonstrated.

Acknowledgements

In conclusion, the author would like to t thankhe Bradford Institute of Technology for providing the facilities for this work, and also all those who have provided helpful encouragement and criticism.

Two-Element Aerial Array

USE OF 'MULTI-FREQUENCY' CARRIER WAVEFORM TO IMPROVE DIRECTIVITY

By V. G. Welsby, Ph.D., M.I.E.E., M.Brit.I.R.E.*

SUMMARY. *A 2-element multiplicative aerial array, used in conjunction with signals whose carrier waveform is composed of the combination of a finite number of sinusoidal components, can be made to give a directional pattern which is comparable with that of a multi-element additive array. This leads to a suggested electronic beam-scanning system which may be simpler than the equivalent one using an additive array. The simplification is obtained at the expense of a somewhat poorer signal/noise ratio.*

While the idea of obtaining improved directional properties by the use of wide-band signals is not new, there is still considerable scope for the development of specialized systems combining particular types of aerial array with suitably-chosen signal waveforms. This applies particularly to systems used for electronic scanning where considerable practical advantages result from any device which reduces the complexity of the electronic circuitry required.

The system to be described can best be explained by means of an example.

Explanation of Method

Consider first a four-element additive aerial array composed of 'point' elements, that is, ones whose dimensions are small compared with a wavelength. Its directional function is

$$D(p) = \frac{\sin 4p}{4 \sin p}$$

which can be written in the alternative form

$$D(p) = \tfrac{1}{2}(\cos 3p + \cos p)$$

This represents the amplitude of the output carrier, at a single frequency ω, as a function of $p = \frac{\omega d}{2c} \sin \theta$ (where d is element spacing in wavelengths and c is the velocity of propagation in the medium);

$$\text{i.e., } D(p) = \tfrac{1}{2}\left[\cos\left(\frac{3\omega d}{2c}\sin\theta\right) + \cos\left(\frac{\omega d}{2c}\sin\theta\right)\right]$$

Now suppose that the two inner elements are removed, leaving a two-element array with spacing $3d$. If the outputs from the two elements are multiplied together and smoothed, the 'd.c.' output will be proportional to $\cos 6p$. Suppose that the signal frequency is replaced by the sum of two equal components at frequencies ω and $\omega/3$ respectively. The output will now be proportional to $\frac{1}{2}(\cos 6p + \cos 2p) = (\sin 8p)/(4 \sin 2p)$. Figs. 1 and 2 show the form of this function, compared with that of the directional function of an eight-element additive array with the same spacing d between the elements as that between the elements of the original four-element array. Note that, since the multiplicative output is phase-sensitive, the first 'diffraction secondary' peak at $p = \pi$ is negative and can be completely suppressed by rectification of the output.

We thus have a two-element array having a directional response comparable with that of an eight-element array which is nearly twice as long. This result has, of course, been achieved at the expense of a reduction in signal/noise ratio and it also requires transducers (or combinations of transducers) capable of responding simultaneously to the two signal components. The signal/noise performance of the proposed system will be considered later where it will be shown that it may be

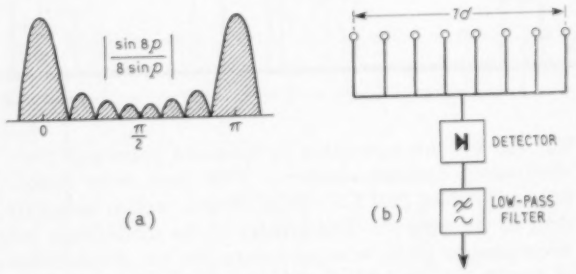


Fig. 1. *(a) Directional pattern of 8-element additive array (single signal frequency ω); (b) block schematic of system*

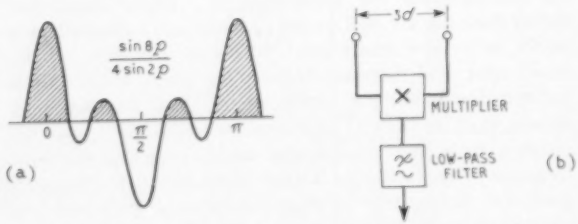


Fig. 2. *(a) Directional pattern of 2-element multiplicative array (complex signal containing ω and $\omega/3$); (b) block schematic of system*

Figs. 1 and 2 *show how a 2-element multiplicative array, in conjunction with a signal comprising n frequency bands ($n = 2$ in this example) may have a directional pattern equivalent to that of an additive array with 4 n elements*

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little better than that of a single element; in other words, the directional pattern of an eight-element array has been obtained but without the improvement in signal/noise ratio which would be produced by a real eight-element additive array. On the other hand, there may be circumstances where this apparent disadvantage may be outweighed by other considerations.

Generalized Theory

It has been shown, by means of a particular example, that a two-element multiplicative array can be given a desirable directional response by the use of a signal waveform which consists of a combination of two sinusoidal components at different frequencies. A more general investigation will now be made of the relationship between directional function and signal waveform for two-element arrays of this type.

Assuming a source sufficiently distant for the wavefronts in the medium to be regarded as plane, the time waveforms of the signals picked up by the two elements, spaced d apart,† will be identical except for a time displacement τ, where $\tau = (d/c \sin \theta)$. If the signal waveform is represented by $f(t)$, the smoothed output of the multiplier will be proportional to the average of the product $f(t) \cdot f(t-\tau)$ over a period which is determined by the bandwidth of the low-pass filter following the multiplier. If this bandwidth is made smaller and smaller so that the averaging time increases indefinitely, the smoothed output will become more and more nearly proportional to the auto-correlation function of $f(t)$ in terms of τ. (In general, it will not necessarily ever become identical with the auto-correlation function because, by definition, this involves integration over all time.)

The Wiener-Khintchine theorem states that the auto-correlation function of any time waveform is equal to the Fourier transform of its power/frequency spectrum. It follows that the directional function of the two-element multiplicative array, expressed as a function of τ, will be given by the Fourier transform of the power spectrum of the signal, provided the integration time is sufficiently large; i.e., provided the latter is chosen so that any increase in the integration time would not lead to any sensible change in the form of the output, expressed as a function of τ.

Now suppose that the particular waveform used is the sum of a finite number of sinusoidal components. When multiplication takes place, various sum and difference frequency components will be formed the lowest of which, apart from the wanted 'd.c.' product, will be equal to the smallest difference to be found between any pair of signal components. Thus, provided the cut-off frequency of the low-pass filter following the multiplier does not exceed this minimum frequency difference, the output will contain only the 'd.c.' component and will be independent of time. (Note that we are not concerned here with the relatively slow time-variation of the 'd.c.' component due to modulation of the complex carrier; as far as directional performance is concerned, the signal may be assumed to have a constant amplitude.)

† Note that, from here on, d is used to denote the actual separation of the elements of the 2-element array.

Under these conditions the output, as a function of τ, will be truly proportional to the Fourier transform of the power spectrum of the signal. Since the latter has been assumed to be made up of a finite number of components, its spectrum takes the form of a series of discrete lines.

An exactly analogous relationship exists between the directional function of an additive array of point elements and the 'sensitivity function' representing the positions and relative sensitivities of the elements.

This means that, in order to obtain the directional function of the two-element multiplicative array, the spectral lines of its signal waveform (including the reflected portion in the negative frequency region) can be imagined as representing, in magnitude and position, the elements of an equivalent multi-element additive array. There is however one important difference; unlike the additive array, the multiplicative array is phase sensitive, with the result that the negative half of the directional pattern can be removed by simple rectification of the output. An interesting point here is that the full advantage of rectification appears only for even numbers of elements since the first diffraction secondary peak is then negative and can be eliminated. This immediately doubles the effective range of τ which can be covered without ambiguity, thus virtually doubling the number of elements in the equivalent additive array. This point is of considerable significance in connection with the problem of electronic scanning which will be considered again later.

It should be mentioned here that, by definition, the directional pattern relates to the relative output of the system expressed as a function of a parameter giving the bearing of a single source. As pointed out in Ref. 1, the response of any multiplicative system to simultaneous signals from two or more sources on different bearings cannot be obtained by simple superposition. Furthermore, the convenient suppression of the diffraction secondary peaks by rectification, as shown in Fig. 2, cannot be used if resolution of simultaneous signals on different bearings is required. Under these conditions, the useful range of ϕ would be limited to $\pm \pi/2$ instead of $\pm \pi$ so that the effective number of elements in this case would not be doubled. This difficulty does not however necessarily preclude the use of the full range of $\pm \pi$ for an electronically-scanned pulsed sonar or radar system, since ambiguity will only occur if the ranges of two or more targets happen to coincide so that the echoes are returned within the same bearing scan.

To obtain an array equivalent to one with an even number of equi-spaced elements, the signal waveform must consist of equal components at frequencies proportional to 1, 3, 5, etc. It has been shown that two components proportional to 1, 3 give a pattern equivalent to that of an additive array with eight elements. In general, a signal waveform with n components spaced proportionally to 1, 3, 5, etc., will give a result, for the two-element multiplicative array, corresponding to an additive array with $4n$ elements.

It is worth noting that the relative phases of the components of the signal waveform are not important; in fact, small deviations from the nominal frequencies could be tolerated since these merely denote slight

errors in the positions of the elements forming the equivalent additive array. In the same way, any amplitude deviations would have the same effect as errors in the sensitivities of the elements of the equivalent array.

In a practical application of this system, the signal components would probably not be transmitted directly but would be shifted up in frequency, appearing in the transmission medium as equi-spaced components lying within a relatively narrow frequency band. The corresponding frequency shift would be carried out at the receiving end so that the signal frequencies again become a series of odd multiples of the lowest component.

Signal/Noise Performance

Suppose that each signal component is accompanied by a band of noise of width $\Delta\omega$; this being the bandwidth necessary to pass the amplitude modulation which will eventually be applied to the complex signal carrier. On multiplication, the lowest cross-multiplication product which can occur will have a frequency of $2\omega - \Delta\omega$. It is assumed that this will be suppressed by the low-pass filter following the multiplier. The 'd.c.' output will thus contain only the sum of the direct products of signal and noise components lying within the respective frequency bands. Consider, for example, the band centred at ω. Assuming that the noises in this band are uncorrelated at the inputs to the multiplier, it is shown in Ref. 1 that the signal/noise ratio after the multiplier will be equal to

$$\frac{R_1 R_2}{\sqrt{\tfrac{1}{2}(1 + R_1^2 + R_2^2)}}$$

where R_1 and R_2 are the respective signal/noise ratios at the multiplier inputs.

If $R_1 = R_2 = R$, this reduces to $\dfrac{R^2}{\sqrt{R^2 + \tfrac{1}{2}}}$

which can be taken as approximately equal to R provided R is greater than about 2.

In fact, the noises picked up from the medium by the two transducers will not be completely uncorrelated, so that the signal/noise ratio is worse than that given above. It can be seen from Refs. 1 and 2 that the correlation factor (assuming point elements in the array and an isotropic random distribution of noise in the medium), tends to a value of

$$\rho\left(\frac{d}{c}\right) = \frac{\sin \omega \dfrac{d}{c}}{\dfrac{\omega d}{c}}$$

for a very narrow bandwidth; i.e., when $\Delta\omega/\omega$ is small. Under these conditions, the correlation is zero when d is any integral multiple of the wavelength in the medium at frequency ω and, although ρ is finite for all other values of d, it cannot exceed $c/\omega d$ and soon becomes negligibly small as d is increased beyond a few wavelengths. A somewhat similar relationship is obtained even when the elements have finite dimensions[1]. A more exact form of the expression for the correlation factor, which applies for all values of $\Delta\omega/\omega$, is

$$\rho\left(\frac{d}{c}\right) = \frac{\mathrm{Si}\left(\dfrac{\omega_2 d}{c}\right) - \mathrm{Si}\left(\dfrac{\omega_1 d}{c}\right)}{\left(\dfrac{\omega_2 d}{c} - \dfrac{\omega_1 d}{c}\right)}$$

where Si ( ) is the sine-integral function and ω_2 and ω_1 represent $\omega + \frac{1}{2}\Delta\omega$ and $\omega - \frac{1}{2}\Delta\omega$ respectively. It is found that the envelope curve of this expression falls even more rapidly than that of $(\sin \omega d/c)/(\omega d/c)$ as d is increased.

It can be concluded that the signal/noise ratio due to one frequency band, at the output of the multiplier, can never exceed that obtained for a single transducer channel and will, in practice, be roughly equal to it.

The same argument will apply to all the signal frequency bands of the system, each of which will produce an output signal/noise ratio equal to that of a single transducer operating in that particular band. Since all the cross-products of different frequency bands will be suppressed by the low-pass filter, the total output noise waveform will simply be the sum of all the waveforms resulting from the multiplication of noises in the individual frequency bands.

The question as to whether there is any correlation between these noise contributions cannot be answered immediately; the answer depends on the way in which the noise originates and is propagated through the medium, but any appreciable correlation is unlikely.

For the present purpose it will be assumed, therefore, that there will be little or no correlation between the noise contributions so that they add together on an r.m.s. basis. Under these conditions it is easily seen that, if the signal/noise ratios in the various frequency bands are denoted by R_ω, $R_{3\omega}$, etc., the effective signal/noise ratio will be given by

$$R_{eff} = \sqrt{n}\left(\frac{1}{R_\omega^2} + \frac{1}{R_{3\omega}^2} + \ldots\right)^{-\frac{1}{2}}$$

The effective signal/noise ratio of the complete system will thus always be somewhat better than that of the worst frequency band and, under the optimum conditions of equal signal/noise ratios in all bands, the system will give an overall improvement of signal/noise ratio by a factor equal to $\sqrt{n}$ where n is the number of frequency bands used.

Application of Electronic Scanning to Multi-Frequency Array

It is explained, for example in Ref. 3, how the directional pattern of an array can be made to scan, by purely electronic means, over a sector the extent of which is usually determined by the need to avoid ambiguity by the exclusion of diffraction secondary peaks. One great advantage of a two-element array is that there is only one channel in which a time-varying delay must be introduced in order to obtain the scanning action. (Note that, in previous systems using a single narrow frequency band, it has been possible to avoid the need for a time-varying delay by using instead a time-varying phase-shift.) Since, in the present system, the need for accurate phase matching of a large number of channels is removed, there appears to be no reason why a practicable delay line with a time-varying delay should not be provided using, for example, the voltage-

sensitive capacitance of semiconductor diodes for this purpose. The transducer or combination of transducers forming each element would, of course, need to respond equally to all the signal-frequency components.

A rather more complicated alternative would be to have separate transducers for the various frequency bands and to insert appropriate time-varying phase-shifts into each of the channels feeding one side of the multiplier. This would provide a way of avoiding the necessity for a wideband variable-delay device and, incidentally, for wide-band transducers. Even though more complicated than the basic circuit using the time-varying delay, this alternative scheme would probably be considerably simpler than the equivalent additive system.

Conclusions

A proposed two-element multiplicative aerial system has been described which uses signal waveforms composed of a number of sinusoidal components. Although the signal/noise performance is not as good as that of a multi-element additive aerial array having an equivalent directional pattern, the principle may be used to produce a simplified form of electronic scanning system.

The design problems involved and the results obtained with an experimental model of this system will be described in a later article.

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New Distortion Criterion

PART 2: A CRITERION BASED UPON THE DISTORTION SPECTRUM

By E. R. Wigan, B.Sc.(Eng.), A.M.I.E.E.

(Concluded from p. 137 April issue)

SUMMARY. *This part of the paper is concerned with the relationship between the subjectively estimated 'unpleasantness' and the purely objective cause—the degree of non-linearity distortion. A formula, described as the 'distortion criterion', is set up in terms of the distortion spectrum, and is justified in several different ways. It survives two critical tests, one of which is the deriving of equal criteria from sine-wave signals which have been clipped once only, or twice symmetrically by diodes, thus conforming to the subjective assessment (established in Part 1), that such wave-forms sound equally unpleasant. The form of the criterion brings out the importance of a quantity related to the signal/noise ratio; experimental evidence as to the level at which this ratio becomes significant is, however, inconclusive.*

The first part of this paper dealt with the reactions of listeners to the introduction of non-linearity distortion into short sections of programme, or into short pulses of pure tone, or into certain notes of short melodies built up of such pulses. It has been shown that the greater part of the data conforms to the statement that "the listeners' reactions to distortion (the 'unpleasantness' of the sound) is conditioned by the time-rate-of-deviation of the signal from normality"—a phrase which will be referred to here as the 'deviation-rate hypothesis'. Among other things the hypothesis predicts (as was confirmed by experiment) that the unpleasantness of a given degree of distortion grows as the frequency rises, and (in certain circumstances) as the signal gets louder.

In this part of the article an attempt is made to correlate the findings of Part 1 with the predictions which will be made from a formula, computed from the distortion spectrum, and described here as the 'distortion criterion'.

The subjective data of the Section headed "The Audience Reaction to Distorted Programme Material" cannot be made use of for, unlike the rest of Part 1, which was concerned with simple sounds which were adjusted to be *equally* unpleasant, that section dealt with complex sounds and with *degrees* of unpleasantness. Equality of unpleasantness can be correlated—as will be demonstrated—with equality of criteria but, to show any kind of correlation between the degree of unpleasantness and the magnitude of the criterion, would involve from the outset a subsidiary investigation of subjective 'scales of unpleasantness'. A solution to this problem has been attempted but the investigation is incomplete and will not be reported here.

In this part of the article it is shown that simple

sounds distorted so as to have radically different tone colour, and sometimes limited to different bandwidths, have approximately the same numerical distortion criterion when they sound equally unpleasant. The agreement is of course not precise but, in the worst cases, usually associated with sounds of radically contrasted tone-colour, the discrepancy in criteria is not more than 15%, and in the best, 3 to 4 %.

Two features of the following analysis must first be disposed of: for purely practical reasons the criterion had to be derived from the *line spectrum* which represented the distortion; this implies a steady signal. *Pulses* of tone, on the other hand, were used when the subjective data were measured. In addition, an arbitrary but necessary decision had to be made in setting up the spectrum for measurement. It has been pointed out earlier that the distortion will rise to a maximum at the crest of the signal pulse, and in an earlier section ("Apparatus and Test Techniques", Part 1) it has been mentioned that a steady signal exactly representing the distortion existing at this epoch of the pulse can be provided by the pulse-generator; the spectrum which will be under consideration throughout this part of the paper, and in the computation of the criterion, is measured from this steady signal.

The criterion was formulated only after rejecting a number of alternatives. The derivation is unimportant and a detailed description would be tedious, the formula [Equ. (1)] is therefore set down without preamble. Its validity is then tested in what follows.

The Distortion Criterion

The proposed criterion C_t is defined as:

$$C_t = \sum_{2}^{n} n^2(p_n - t) \text{ . . . (per cent)} \quad .. \quad .. \quad (1)$$

for $p_n - t$ positive.

where n = the harmonic number; ($n = 1$ implies the fundamental);

p_n = the percentage amplitude of the nth harmonic;

and t = that harmonic percentage which is just audible under the conditions of the experiment.

The definitions of two of these symbols need amplification; first, the percentage p_n is to be taken in terms of the amplitude of the original pure sine wave from which the distorted signal was derived. If the distortion is small there is no significant difference between this and the amplitude of the fundamental component *present in the distorted signal* but, in some cases, the difference is important (See App. 5).

Secondly, from the definition given above, it is obvious that the value of t must vary in some way with the threshold of hearing of the individual listener and, in any case, to represent the normal rapid increase of hearing loss at high frequency, t must eventually become very large as frequency rises. For the purposes of this Article, these factors have had to be ignored and t has been treated as quite independent of either n or the frequency of the test signal. The justification for this drastic simplification is doubtful, but the necessity for it will appear later.

Thirdly, t will fall if the test signal gets louder. The point does not arise in the present discussion because all the signals were presented at the same loudness but should an attempt be made to apply C_t to programme of continually fluctuating level, some allowance must be made for this variation in level.

Computing the Criterion

All spectra discussed apply to a sine wave which has been 'clipped' by a diode; the computed 'weighted' spectrum of the signal which results when the diode is *ideal* is shown in Fig. 14 (a) for all values* of the quantity d/T. This and the fraction q by which the wave has been de-limited are defined in Fig. 14 (b).

The curves in Fig. 14 (a) carry the harmonic number n of Equ. (1). The Y-co-ordinate is n^2p_n. To calculate the criterion C_t the quantity d/T must be known; the terms of the weighted spectrum are read off the various curves at that point on the diagram. Since the diagram ignores t, the quantity n^2t has to be subtracted from each ordinate *before making the summation called for in Equ.* (1). (The sign of the ordinate is ignored in the summation but, for other reasons, retained in following diagrams.)

The resulting total is the criterion appropriate to the wave-form in its original; i.e., undiluted, state; but if the distortion has been diluted the ordinates have to be reduced, in the proportion $S/10$, *before* n^2t is subtracted. In practice t is likely to lie between 0·1—0·5%, in which case the numerical value of the criterion will fall much more rapidly than would be expected in view of the degree of dilution.

* In Fig. 14 (a) a limit is set at $d/T = 0.5$ because, in practice, no distortion condition can occur which corresponds to higher values.

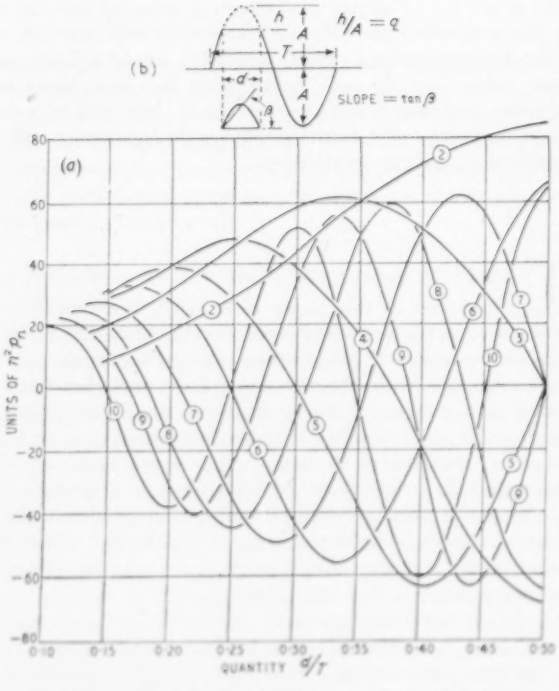


Fig. 14. *Weighted spectra of ideally-clipped sine wave. The parameter is the harmonic number ($n = 1$ implies the fundamental)*

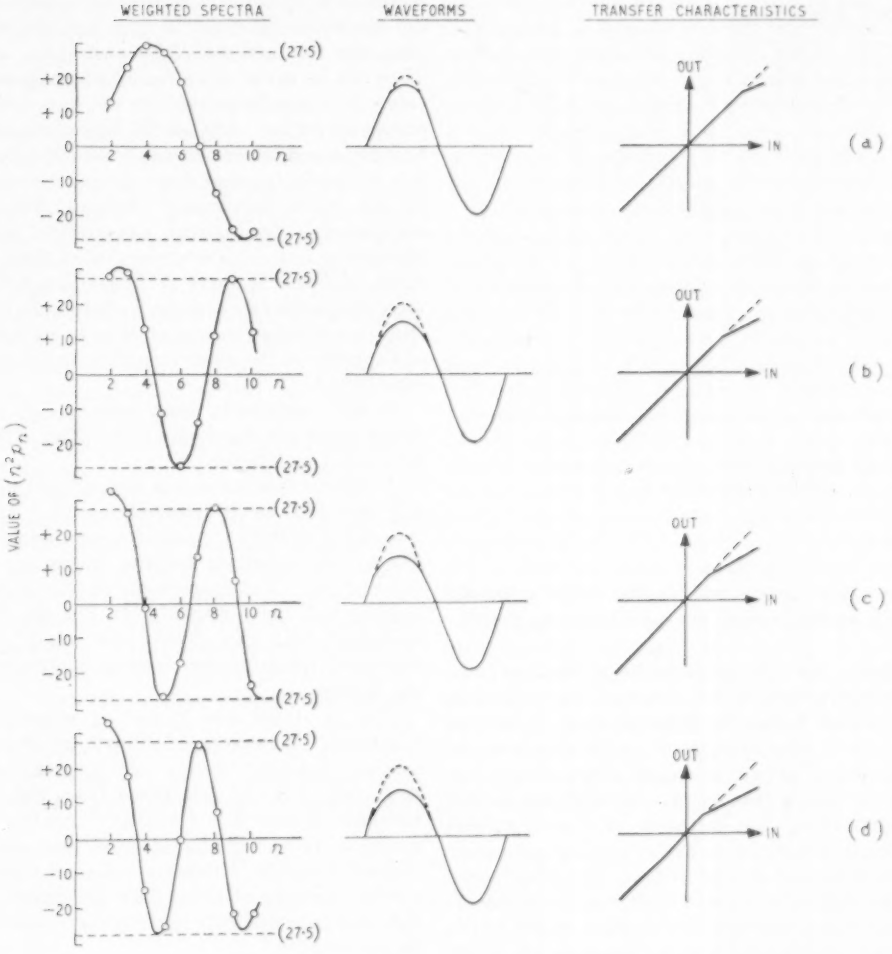


Fig. 15. *Four equally unpleasant sounds. The weighted spectra and the transfer characteristics of the clipped waves of Fig. 4*

In the following discussion t is occasionally written as zero in Equ. (1), to simplify the analysis. The criterion is then written C_0 in place of C_t, it being understood that in both cases all terms up to the tenth [the highest tabulated in Fig. 14 (a)] are included.

The Weighted Spectra of Four Equally-Unpleasant Sounds

Wave-forms dealt with in Fig. 4 reappear in Fig. 15 together with their weighted spectra, which have been derived from Fig. 14 (a) (the distorting diode is, therefore, taken as idealized) because detailed measurements were not considered important when the original subjective tests were made. It can be seen at once that the weighted spectra have at least one feature in common, namely the limits (about 27 units of n^2p_n) between which their ordinates oscillate. On working out the value of C_0 it is found that the criteria are almost identical, varying from the mean by less than 3%. This correlates with the subjective judgments of the earlier section ("Examination of the Difference Signal", Part 1) that they sounded equally unpleasant.

From the spectra in Fig. 15, it can be deduced that, had the bandwidth of the transmission system been limited by, for instance, a filter or by the loudspeaker system having a sharply defined cut-off frequency, the criterion would be reduced almost in proportion. This situation is imitated in Fig. 16 where the summation of weighted terms has been carried up to various values of n. As might be expected, the four values of criterion do not agree well when the bandwidth includes only a few terms, but it can be seen that, in general, the criterion C_0 varies directly as bandwidth and that the four criteria become more nearly alike as the bandwidth increases.

The Influence of the Term t in the Criterion Equation

Significance of "t"

The terms plotted in Fig. 15 and summed in Fig. 16 are of the form n^2p_n. Equ. (1) uses $[n^2(p_n-t)]$, but no guide is given to the appropriate value of t. Simple substitution of fixed values of t (say, 0·1, 0·2, 0·3 . . . 0·5%), shows that the equality of the four criteria in

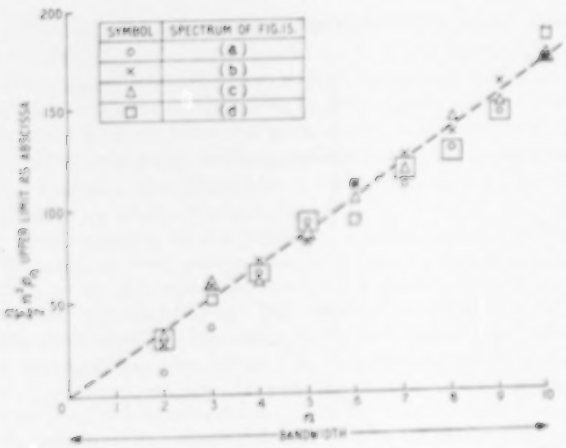


Fig. 16. Four equally unpleasant sounds. The effect of bandwidth limitation upon the spectra of Fig. 15

Fig. 15 is virtually unaffected by the value selected, although the magnitude of all the criteria shrink rapidly as t grows.

In this respect, Equ. (1) conforms to practical experience, for it was observed in the course of the subjective experiments of Part 1 that the ambient room noise had little effect upon the *equality* of unpleasantness of distorted signals, although it made them sound less objectionable. The inclusion of t somewhere in the equation representing the criterion is thus justified, but to establish its proper value in any given experiment, and the place it should occupy in Equ. (1), is not easy.

The primary difficulty is that ambient noise represents only one factor contributing to t. Another, which in the quiet situations used in the present experiments is the more important, is the masking-effect produced by the strong fundamental component of the test pulse—an effect which cannot be predicted with any great precision even if the sound level and all other relevant factors are known. Lazenby[3] has reviewed the position making reference to, among others, von Braunmuhl's work[2], which has direct application to the subject under discussion here. Lazenby and others noted[4] large discrepancies between the degree of masking which would be predicted from published data relating to the generation of the so-called 'aural harmonics' (which are supposed to produce the masking effect), and the actual degree of masking deduced from purely subjective experiment. Discrepancies of 10 dB are mentioned by Lazenby, and von Gassler is not prepared to claim that his theoretically-deduced data agrees with experiment within better than $\pm$ 10 dB (a range of 20 dB). It is apparent, therefore, that until the masking phenomenon is more exactly understood, the value of t must be derived experimentally in each case.

An Experimental Determination of t

Fig. 17 illustrates one method of fitting a value of t to a set of data. Having first chosen for comparison a pair of spectra A and B of *different bandwidth*, and having established by subjective test the degree of dilution of the distortion necessary to give equality of unpleasantness, the two criteria $(C_t)_A$ and $(C_t)_B$ are computed. They can be made equal by substituting the appropriate value of t simultaneously in the two criteria. In the previous section ("The Weighted Spectra of Four Equally-Unpleasant Sounds") this process failed; here it is successful because there are more terms to be added in the wider spectrum. Because these terms have successively diminishing amplitudes, quite a small increase of t can obliterate several of them, and, because those obliterated have a heavy frequency-weighting, their disappearance reduces C_t drastically; the narrower spectrum is much less sensitive to t. In this way, a value of t specific to the given experimental conditions can be established.

In the experiment dealt with in Fig. 17, the appropriate value of t was found to be 0·3%. The experiment proceeded as follows.

A 500-c/s sine wave was clipped so that the quantity d/T [see Fig. 14 (b)] was about 0·4. Two test-signals A and B of different bandwidth were selected by low-pass filters; the weighted spectra are shown in Fig. 17 at (a) and (b). The spectra are shown at the stage when neither had been diluted. In (b) the full line is the measured and the dotted the ideal spectrum; the measured spectrum was used in constructing the rest of the figures.

The A signal was presented without any dilution whatever; B was then diluted as required and the scoring function shown at (c) was experimentally deduced. Fig. 17 (c) differs from Fig. 2: the X-coordinate is not the quantity S but the value of the criterion $(C_t)_B$ calculated from the spectrum of the diluted B signal. Arbitrary values of t are chosen and a similar number of score lines are drawn. The scoring function indicates that in this experiment t should be put at about 0·3%.

The method of arriving at this figure is as follows. The spectrum of the A signal was fixed throughout the experiment and the value of the criterion $(C_t)_A$ was calculated for a range of values of t as shown at (c). The B signal had to be treated differently, for the distortion was diluted by four different amounts during the experiment so that the spectral amplitude was reduced to 50, 60, 70 and 80% of that shown in (b), and to each degree of dilution a different value of the criterion applied.

The four values of t used in the diagram (c) were selected as best illustrating the influence of t. The score line based upon $t = 0{\cdot}3$ passes through the particular triangle-and-circle which defines the value of $(C_t)_A$ calculated also from $t = 0{\cdot}3$ (the arrows show which of the other A-criteria are associated with each of the other score lines). At this value of t therefore the criteria $(C_t)_A$ and $(C_t)_B$ agree precisely. At the same time the intersection with the 50% score line proves that the two signals sounded equally unpleasant.

The audience scores do not fit well to any smooth curves because observers were upset by differences of tone-colour. In this experiment, and in others, the value of t which had to be inserted in Equ. (1) was higher (possibly 10 dB higher) than the observers' hearing threshold as measured in the ambient noise. On this

occasion the effect of the ambient noise alone was measured and found to be equivalent to about $t = 0{\cdot}1\%$ which is to be compared with the effective value of $t = 0{\cdot}3\%$ just deduced.

The Unpleasantness of Singly- and Doubly-Clipped Waves: the Resolution of an Anomaly

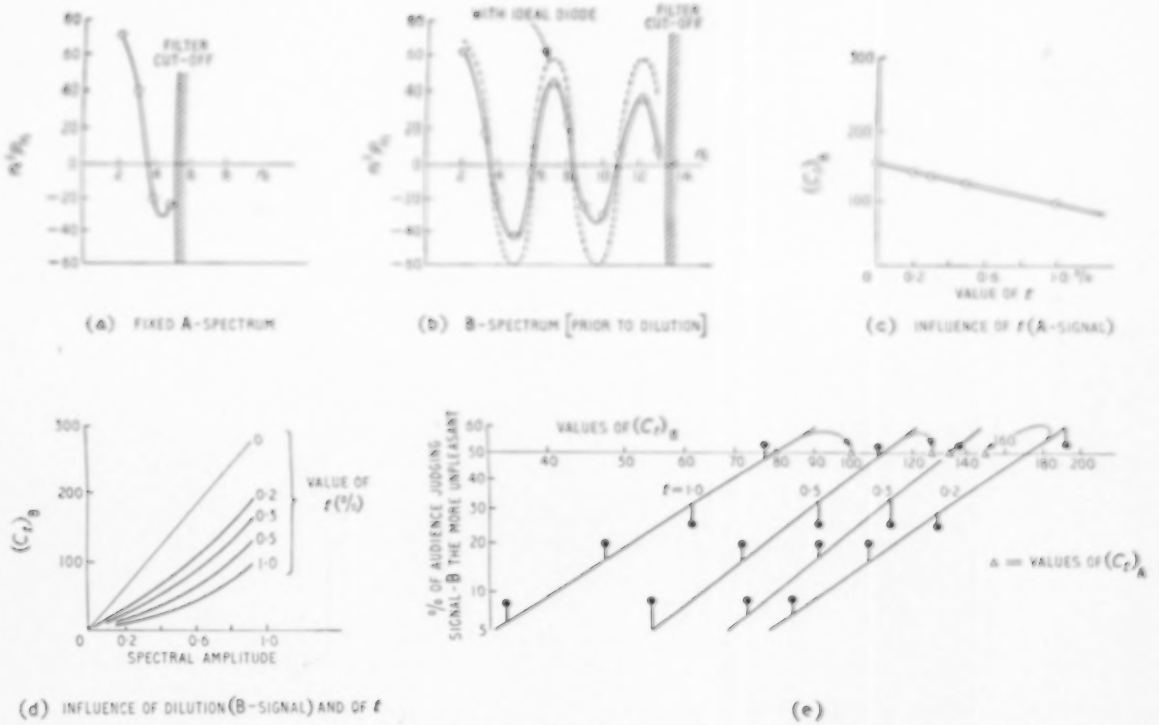
A more direct test of Equ. (1), and one which avoids the controversial quantity t, is found in wave-forms which have been doubly-clipped (see earlier section: "A Crucial Experiment", Part 1); the spectrum which is generated by this operation is so highly characteristic that the value of t can be ignored, at least in discussing broad effects. All the various signals, the spectra and transfer characteristics of which are given in Fig. 18, were found to be as unpleasant as the first signal (a) which, alone, was singly-clipped.

All the spectra (b) to (g), in spite of wide differences of pattern, had criteria C_θ which lay within $\pm 15\%$ of the criterion calculated from the spectrum of (a). In view of the considerable difficulty of judging sounds which differed in tone-colour as markedly as these did, and the use of the highly-artificial criterion C_θ in place of the *more realistic* C_t, the agreement between the subjective and objective evidence can be considered good, and the outcome of this series of experiments a satisfactory and severe test of the validity of the criterion.

It can be seen from the transfer characteristics that Fig. 18 (a) and (d) apply respectively to singly- and doubly-clipped waves in which the degree of distortion (determined by the values of q and S) is the same; it is because of this that the associated weighted spectra are quite regular, while none of the others are. The explanation is given in App. 4; the even terms of the (d)-spectrum are all zero, while the odd ones have precisely twice the amplitude of those in (a). On this basis, therefore, it is obvious that the criteria for the spectra of singly- and doubly-clipped waves must agree very closely, the lack of the even terms being, in the long run, compensated by the doubled amplitude of the remaining odd ones.

If the spectra were limited to a few terms, the two criteria would not be precisely equal—an infinite series would be required for this to be so—and it is difficult to demonstrate this equality from the extremely limited data of Fig. 14 which runs no farther than $n = 10$. Moreover, it is not possible to use more than eight of these terms ($n = 2$ to $n = 9$), because a fair comparison can be assured only if there are twice as many terms in the singly-clipped spectrum as in the doubly-clipped. On this basis, Fig. 19 has been constructed; it shows that the two criteria are roughly equal. For reasons to be mentioned below the X-axis of the figure has been made linear in terms of the slope of the difference signal, as defined at (c). The quantity d/T is defined at (b). If d/T is made 0·5 the doubly-clipped wave vanishes; this is why curve (2) falls to zero.

Turning now to the subjective evidence: equality of unpleasantness of singly- and doubly-clipped waves has been confirmed experimentally as mentioned in Part 1 up to the point at which $d/T = 0{\cdot}3$, but to go further is difficult because of the radical difference of tone-colour of the sounds being compared. Fig. 19 agrees; up to this point curves (1) and (2) differ only slightly. The dotted line in the figure clears up an earlier diffi-



(a) FIXED A-SPECTRUM (b) B-SPECTRUM [PRIOR TO DILUTION] (c) INFLUENCE OF t (A-SIGNAL)

(d) INFLUENCE OF DILUTION (B-SIGNAL) AND OF t (e)

Fig. 17. *Experimental determination of the factor t; the value of t is deducible because the spectra compared have unequal bandwidth. This data fits $t = 0{\cdot}3\%$*

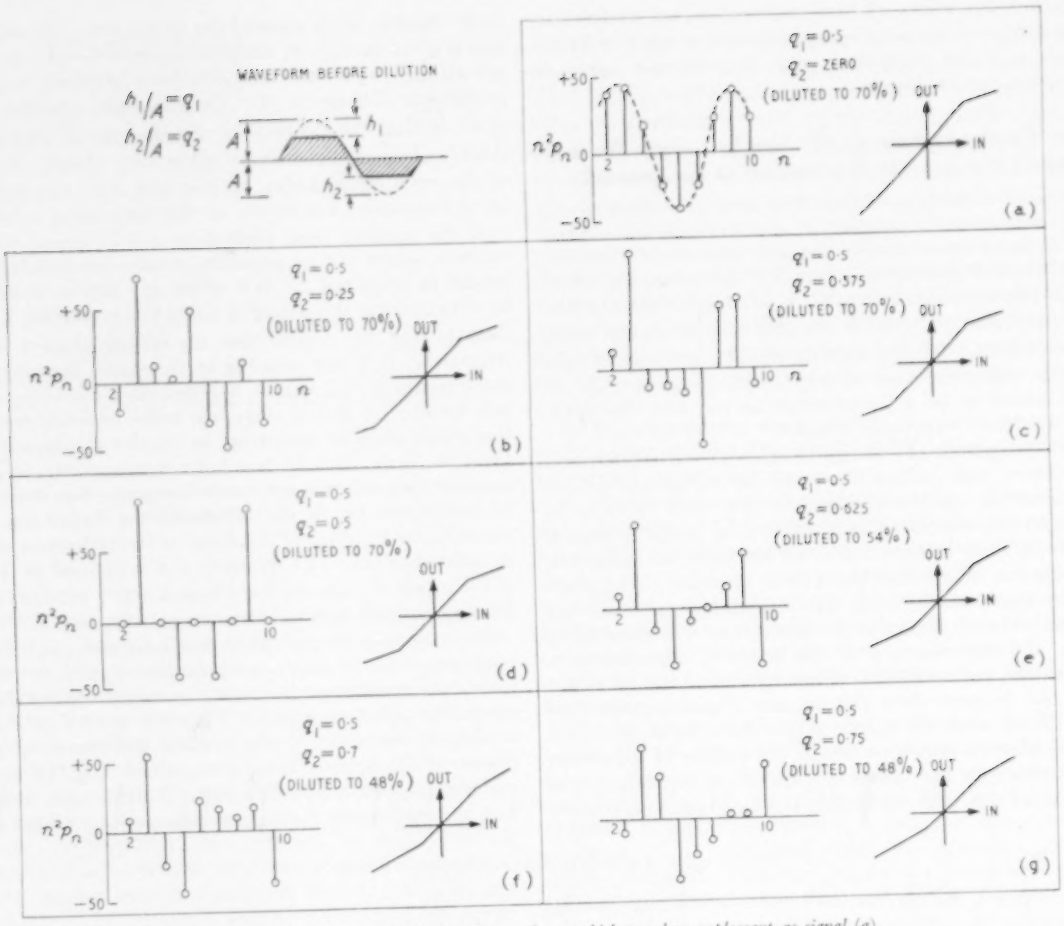


Fig. 18. *Six different doubly-clipped waveforms which sound as unpleasant as signal (a)*

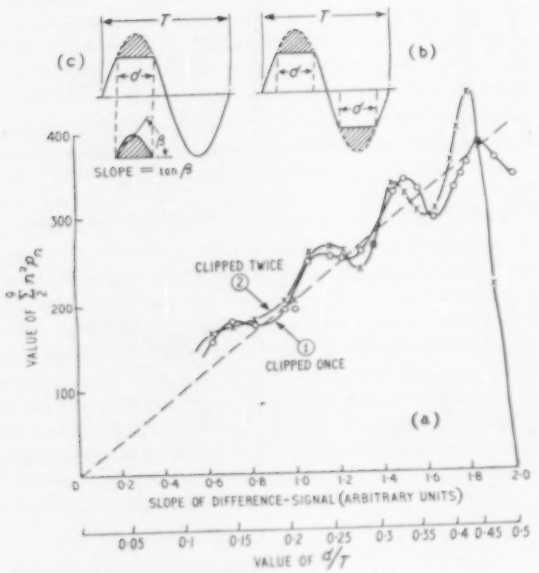


Fig. 19. *Relation between distortion criteria of singly- and doubly-clipped sine waves. Criteria for singly- and doubly- (symmetrically) clipped waves are shown to be similar and roughly proportional to the slope of the difference-signal*

culty, it shows that C_0, whether for a singly- or doubly-clipped signal, tends to vary as the slope of the difference signal. We deduce that the anomaly of the earlier section ("A Crucial Experiment") was an artificial one —unpleasantness is indeed related, as was shown earlier in Fig. 4, to the slope of the difference-signal, but it is *not* concerned with the rate of occurrence of distortion as had previously been assumed.

Discussion of Parts 1 and 2

From the extensive evidence available the bare minimum has been selected to support the thesis of this paper. The analysis presented here can be criticized as being incomplete not only on this account, but also on a quite different basis, namely:

That most of it deals only with the distortion of simple sine waves; that the distortion is limited to a 'clipping' operation; that the pulse test frequencies lie in a region between 250 and 1,500 c/s—the greater part in the even narrower region between 250 and 1,000 c/s; that a critical factor t in the criterion equation is incompletely defined; that nothing is said about the significance of the *magnitude* of the criterion (as distinct from its comparative significance); and, finally, that no way is

suggested by which the criterion can predict the 'unpleasantness' of distorted programme signals. There are, however, good reasons for the failure to close these gaps in the argument.

Dealing with the last two points first, the extension of the analysis from pulses to complex programme signals has already been attempted, but is too elaborate to deal with here. In any case, it appears that a new subjective judgment technique will have to be devised in order to obtain reliable conclusions against which to test the predictions made from the criterion, and this has not yet been done.

As to the criterion magnitude: a rough approximation, based upon limited experimental evidence, is that if C_t is in the region 10/20, distortion will be on the verge of being noticed, and if greater than 100/200 it will be highly objectionable[5].

Reverting for a moment to the subjective tests of an earlier section ("The Audience Reaction to Distorted Programme Material"): it is more than likely that the reader has already registered the objection that the test-conditions were highly artificial, and that judgments of long sections of programme, some not appreciably distorted, some grossly and some only partially, would lead to radically different conclusions. This is quite probable. The justification for this earlier section, on the other hand, is that it shows that several of the subjective phenomena revealed by the pulse and the melody tests reappeared under programme conditions (however unrealistic): it was on this count that these phenomena could properly be classed as 'behaviour-patterns'. This being established it is not unreasonable to suspect that they will reappear, although modified in a way at present unpredictable, when some new subjective test technique is devised to deal with long-period programme assessment.

The justification for the close limitation of the test conditions criticized above is largely on the same grounds: the frequency-range was de-limited because in the chosen region the most striking subjective effects were found; the diode was retained as a distorting device after the need had been realized for a very wide distortion-spectrum with which to test the validity of the distortion criterion and because, with the help of simple filter-assemblies, any of the more elementary forms of distortion could readily be imitated; as explained in "Apparatus and Test Techniques", Part 1, it has other justifications.

The lack of precise information on the factor t is unfortunate. The inadequacy of the highly complicated method of arriving at it as set out in "The Influence of the Term t in the Criterion Equation" need not be laboured, and some quite different lines of attack will have to be devised; several are immediately obvious and some have been tentatively investigated. For instance, bandwidth-limited noise may be added to the signal and its effect upon the subjective judgments and upon the criterion compared; the value of t appropriate to the criterion is then found to vary consistently with the noise level until the noise gets too weak to compete with the masking effect produced by the fundamental tone of the test signal; from such tests the masking-effect can be very approximately deduced. Again, if diagrams such as Fig. 2 are plotted for the case in which the distortion of the two signals compared is very small, the scoring function is curved—particularly if the distortion (although of very small amplitude) has a large spectral range; from this curvature t may be estimated, but the method is oblique and the accuracy low.

The real significance of t lies not in its possible magnitude, but in its appearing in the criterion at all. Its presence makes evident at once the importance of the signal/noise ratio under which the listening tests are made: the softer the sound the greater the (objective) distortion which will be tolerated and, conversely, the louder the sound the more unpleasant any residual distortion will become.

It is the influence of t combined with that of the frequency-effect which is responsible for the sometimes disastrous outcome of combining an increase of bandwidth with an increase of sound level in sound-reproducing equipment: the higher-order (and in the usual order of events the smaller) components of the distortion spectrum are easily masked by noise; a small increase of signal/noise ratio makes them audible; and, finally, they have a very heavy subjective weighting. All of these factors contribute to the overall unpleasant effect.

The distortion criterion can be validated only by testing it under a wide range of completely different experimental conditions and this has not yet been feasible; nevertheless, the criterion has been shown to make the correct prediction in two groups of critical test cases—the equality of unpleasantness of singly- and doubly-clipped sine waves, and of waves in which the difference signals have similar slopes.

The deviation-rate hypothesis, on the other hand, is readily supported from general experience, at least in so far as it postulates that the unpleasantness of any distortion depends upon the departures from *normality*. Fig. 9 (f) makes the point quite clear—distortion can be recognized but cannot be assessed if the context is strange. The violent changes of tone colour common in modern forms of popular music would provide a similarly unpredictable context, and one which must render this kind of material almost immune to distortion. By contrast, piano programmes, as is well known, provide highly critical test material for, mutatis mutandis, the tone colour of every note is predictable from its predecessor.

The delay phenomena of App. 5 do not fall into the main line of development of the thesis; they have important theoretical implications, but these are best dealt with in that Appendix. A practical point must be made however: Appendix 5 shows that distorted programme-material sounds less objectionable if heard through high-quality headphones, rather than via a loudspeaker. The inference from this is that high-quality programmes cannot be monitored successfully by headphones—no matter how high their quality—if the audience for which the programme is intended receives it by loudspeaker.

Summarizing the situation: the analysis of Part 1 of this paper shows that certain observer-behaviour patterns exist which reveal themselves in artificial and also in real-life situations; Part 2 suggests a numerical criterion which has justified itself not only in showing a high correlation with certain critical subjective data, but also in explaining, in reverse, an anomaly discovered

in Part 1. The criterion seems, on these and other grounds, to be self-consistent to a useful extent, and should be usable to predict the relative unpleasantness of distorted sounds, provided the various test conditions, including the signal/noise ratio, are closely defined. However, until more is known about the quantity t, the predictions may fail if systems of widely different bandwidths are compared. The prediction of the audience reaction to the distortion of programme material must for the present be made with caution, for Part 1 of this Article makes it clear that the reaction can be strongly influenced by a variety of subsidiary circumstances*; the incidence of these will modify any predictions based solely upon the distortion criterion.

It is difficult to find published reports with which to compare the present work, and most of the references given below have bearing only upon the quantity t; however, attention should be drawn to Ref. 6 in which D. E. L. Shorter of the B.B.C. Research Department made an important (though no more than tentative) examination of certain distortion criteria. He developed by trial and error a formula which was adequate to place five radically different transmission systems in the rank order assigned by listeners. However, although his criterion contained, as does Equ. (1), terms in n^2p_n, *the terms were squared before summation*, the total being finally square-rooted. This operation obviously derives from the conventional total-harmonic measurement but has no other specific justification. In an extended investigation of the validity of this criterion, undertaken by the writer, it was found to be not entirely reliable in predicting audience reaction. Part 1 offers an alternative and crucial test—singly and doubly-clipped sine-waves sound equally unpleasant and therefore must bear equal criteria: Shorter's formula fails to meet this condition, giving a ratio of approximately 1·4/1 in favour of the doubly-clipped wave.

The equality of unpleasantness of signals of this kind was, by a fortunate accident, established very early in the progress of the present work: in conjunction with the closely following enunciation of the deviation-rate hypothesis, r.m.s. addition of distortion terms was ruled out, and thus a fruitless investigation of Shorter's criterion was avoided. It is interesting to note, however, that the major difference between that criterion and Equ. (1) lies simply in the method of aggregation of the terms of the formulae. It can now be seen that 0·03% (the value of t adopted arbitrarily by Shorter) is unrealistic in view of the probable values of signal/noise ratio operative during his investigation.

Acknowledgements

This investigation is concerned with human behaviour and this paper could not have been written without the help of many observers. The 30 senior members of B.B.C. staff concerned in the programme tests put on record some thousands of data entries, each of them the outcome of concentrated thought and considered judgment. Over a period of some two years, some 20 members of the B.B.C. Research Department staff provided a much larger mass of data; they took part not only in all the other experiments described here, but also in the many abortive investigations of which no mention is made. Acknowledgements are made for their patience and ready assistance in a task which is always tedious and often mentally exhausting, and for the precision of their work which made it possible to bring some degree of order out of what was at one stage a formless collection of data.

To carry out this investigation a large amount of work was involved in the construction, operation and maintenance of the ancillary technical equipment, and also in the organization and analysis of the subjective experiments. Acknowledgements are made to Messrs. P. Gouldstone, R. L. Deane and R. J. Packer of the B.B.C. Research Department for their assistance in this respect.

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APPENDIX 1
Distortion Generator

(*a*) *General Arrangement*

The input signal is passed through four separate channels [see Fig. 20(a)]; the upper provides a 'clean' version which can be selected by the switch K_2 in its middle position, the two centre channels are passed into the distortion-generator network containing a diode, biased by a voltage V_b, and a series resistance used to reduce the degree of distortion. The diode is isolated by high series resistances to ensure that its response is independent of frequency. The fourth channel is used to generate the difference signal, so that the distortion can be checked.

(*b*) *Means for Varying the Distortion*

The distorted signal is passed to a potentiometer P_1 tapped in ten equal steps. Two separate rotary switches, S_A and S_B select proportions of the distorted signal from P_1 and pass them to the respective 'A' and 'B' contacts of switch K_2.

Ganged contacts on S_A and S_B select *complementary* proportions of 'clean' signal from potentiometer P_2 and feed them also to the 'A' and 'B' contacts of K_2.

In this way, distortion appearing in P_1 is 'diluted' by clean signal in P_2 to an extent depending on the settings of S_A or S_B. The effect of dilution is illustrated in Fig. 20 (b), inset; (for simplicity, V_b is assumed here to be zero and the diode perfect).

Conditions in the first diagram correspond to $S_A = 0$, switch S_A standing on the contact marked '0'; there is no distortion and the transfer characteristic is a straight line as shown.

When $S_A = 10$ [third part of Fig. 20(b)] rectification is complete, and the right-hand part of the transfer characteristic is horizontal. For any intermediate value of S_A the slope of the right-hand end of the characteristic varies in proportion to S. The left-hand side remains unchanged throughout.

The same facility is provided for the 'B' signal so that different and precisely known degrees of distortion can be fed to the loud-

* As, for example, a sudden variation in the scoring of a musical item. The data in Fig. 9(c)., when analysed on the lines adopted for the piano test-piece, revealed a violent transient change of grading (of the order +4 to −4dB) at a point at which the scoring of the military-band music was modified in favour of the string section rather than the brass. (The influence of the associated change in mean frequency of the programme material had already been allowed for in the analysis).

speaker in the 'A' and 'B' test signals by operating the switch K_1. Additional contacts on K_2 change the value of V_b, insert filters in the loudspeaker circuit, or change the test pulse frequency, as required for the experiment in hand.

In the main text, the degree of distortion is defined by the value of S. Thus if $S = 6$, say, it implies that 6/10ths of *whatever distortion the diode circuit may have been arranged to produce* is present in the B signal when it leaves via K_2.

APPENDIX 2

Signal-Pulse Generator

The signal-pulse illustrated in Fig. 1 is generated by the circuit of Fig. 21, operating as follows:

The steady current in the diodes is reversed by switch S_1, passing through zero, as shown at (e), a certain time after S_1 is operated. The delay before the zero is reached is controlled by the potentiometer in the associated RC network.

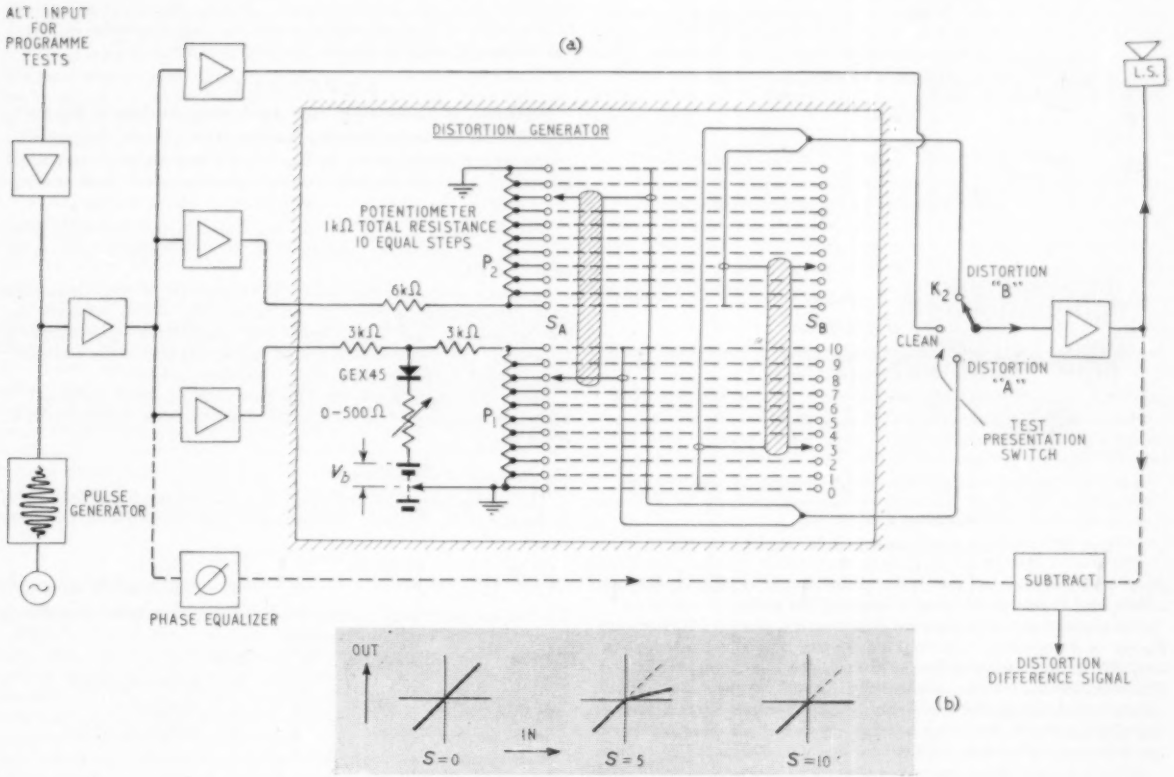


Fig. 20. *Simple circuit to allow dilution of distortion without change of spectral composition*

In Fig. 20(b), V_b is zero; when V_b is relatively large, the transfer-characteristics are as in Fig. 18. A diluted signal has precisely the same spectrum as the undiluted one except that the amplitude of every term is reduced in the proportion $S/10$.

(*c*) *Production of a Signal in which Distortion Percentage is Independent of Signal Amplitude*

Distorted programme signals could not have been analysed if the distortion had varied with instantaneous signal amplitude. The percentage was kept constant as follows:

Suppose that $V_b = 0$, and $S_B = 10$; rectification is perfect and every positive-going element of the signal is suppressed; then it is quite clear that, irrespective of the signal amplitude, the amplitude of the distortion-difference signal will be exactly half that of the input wave. *In terms of the distortion-difference*, the degree of distortion is precisely 50% at all signal levels.

So also the distortion will be 25% when $S = 5$, and so on. In the work of the section headed "The Audience Reaction to Distorted Programme Material" about 250 Ω was added in series with the diode to reduce its efficiency, but the general principle still applied and, on measurement, it was found that the distortion ratio remained substantially constant for any given setting of the dials, irrespective of signal level.

The larger the control current in the diodes the larger the signal current they allow to pass; with control current at zero, diode resistance rises and reduces the signal current, as shown at (c). This signal (from winding 'A' of the transformer) is the reverse of the pulse required, and the desired pulse is formed by subtracting from the signal (c) the steady signal (b) available in winding 'B'. The result shown at (d) is obtained when the magnitude of (b) is set to reduce (c) to zero when the control current is steady.

Pulses produced by this method always have the same proportions (see Fig. 1 of the main text) irrespective of their width T_p which is defined as the width, in milliseconds, measured at half the height of the pulse. With the circuit components shown, T_p is variable from 10 to 65 msec; in most experiments it was 40 to 60 msec.

A low-pass filter is needed in the output circuit to remove the distortion terms generated by the diodes.

In this investigation, the most important feature of the pulse was its peak amplitude; in this circuit, it is highly consistent from pulse to pulse.

The alternating voltage corresponding to the peak of the output pulse can be measured under steady-state conditions by opening switch S_2. The bandwidth of the pulse is small ($T_p > 40$ msec) and the cut-off of the low-pass output filter can be set so that very

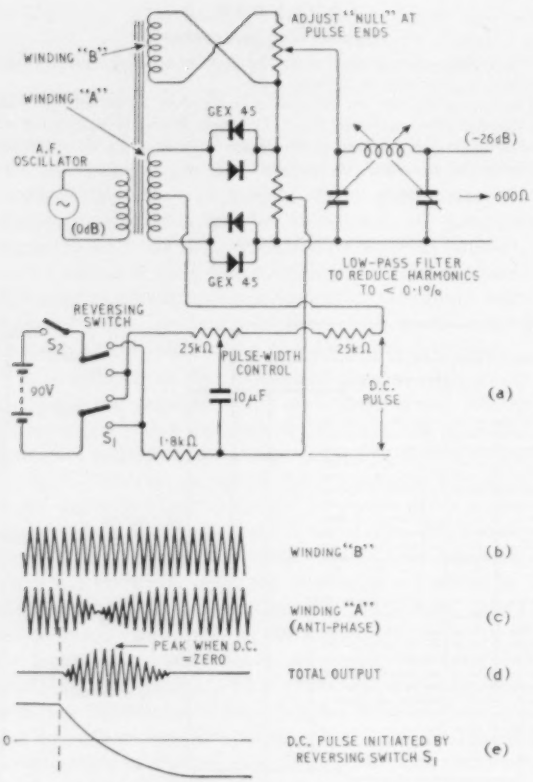


Fig. 21. *Pulse generator; produces pulses shaped as in Fig. 2*

little pulse energy is lost; the steady-state reading will then correspond closely to the peak voltage of the pulse.

For the melody experiments of the section headed "Elementary Forms of Programme Material . . ." the pulse was switched on and off by the voltage delivered from the RC network. This was first rectified and then arranged to bias-off a pair of balanced valves, which controlled the signal level. A small piano keyboard was provided with the contacts necessary to carry out, in sequence, the following operations as any one key was struck:

(a) Cause the previously generated pulse to decay very rapidly to zero so that there was silence for a few milliseconds before the next note sounded;

(b) tune a continuously running RC oscillator to deliver the correct frequency;

(c) operate the reversing switch (S_1 of Fig. 21) to initiate the pulse;

(d) energize the magnet of an automatic line-finder switch, which on the release of the piano key, and just before the next was pressed, stepped forward and put into effect the degree of distortion and/or the signal loudness appropriate to the next note in the melody.

By this means, even when the keys were operated rapidly in sequence, consistently shaped pulses were produced and precisely repeatable degrees of distortion or stress were applied to each note; at the same time the notes were free of any noticeable starting or stopping transients. For the experiments of the Section headed "Elementary Forms of Programme Material. The influence of 'Context' upon the Unpleasantness" (Part 1), the pulses had a half-height-width of 40 msec, and sounded very like the pulse of Fig. 1 but decayed to zero in a finite time of the order of 2 T_p.

APPENDIX 3

Methods of Collecting and Analysing Audience Scores

(a) Electrical Method of Summing Audience Scores (Graded Response)

The arrangement shown in Fig. 22 will cope with six grades of assessment; for this investigation five positions were used marked:

'A'+ 'A'− 'EQUAL' 'B'− 'B'+

A vote registered as 'A'+ meant that sound A was obviously more distorted that sound B; registered as 'A'− it implied, 'observable only with effort'.

By means of the selector switch S_1 the number of votes cast under each classification is read off on meter M_1, which registers the conductance of the loop; each vote contributes approximately 55 μmho. The meter-shunts are set to suit the size of the audience; any number may take part, each with his own key.

Meter M_2 which reads the total votes cast is found essential; it allows the operator to check that all observers have voted, before he reads M_1.

(b) Reduction of Graded Scores

The voting has to be converted to a simple percentage judgment in favour of B. Suppose the scoring in one test with six observers runs:

'A'+	'A'−	'EQUAL'	'B'−	'B'+	(Grade)
1	3	1	1	0	(Number of votes cast)

To find the percentage in favour of B each grade is given a weighting factor, the first being made zero:

'A'+	'A'−	'EQUAL'	'B'−	'B'+	
1	3	1	1	0	Scores recorded
0	1	2	3	4	Weights
0	3	2	3	0	Weighted scores

Total weighted scores = 8.

The maximum possible weighted score is 24; (six observers all scoring 'B'+). The proportional score in favour of 'condition B+' expressed as a percentage p is: $p = (8/24)\ 100 = 33{\cdot}3\%$.

(c) Plotting the Data to obtain the Value of S_{BM}

Fig. 2 is typical of several hundred diagrams used in the preparation of this article. The Y-axis is based on the probability integral

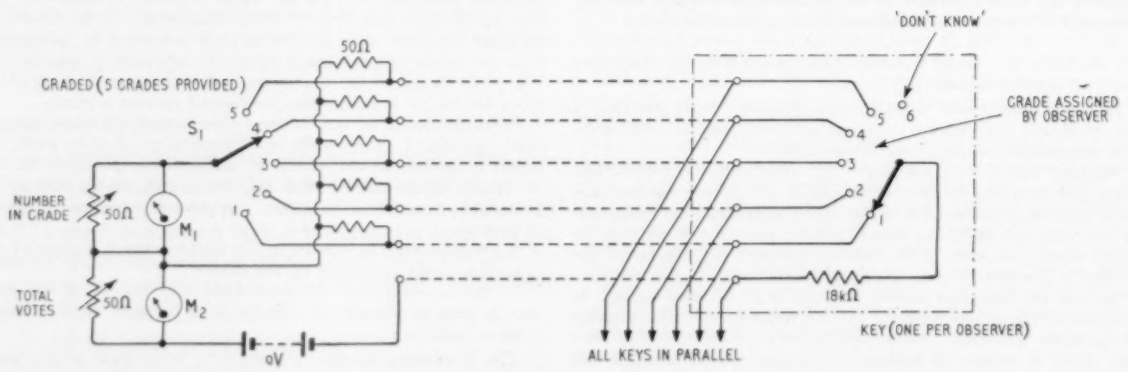


Fig. 22. *Audience-score analyser; switch S_1 at each step shows on meter M_1 the number of observers who voted for the grade indicated. M_2 reads the total number voting*

and the X-axis is logarithmic. Generally a straight line could be fitted to the plotted points. (Curved scoring functions are mentioned in section—"Discussion of Part 1 and 2".)

For conciseness, the ordinates in Fig. 2 are marked: "likelihood of voting in favour of B as the more unpleasant." Strictly, the ordinate should therefore be treated as representing "the best available estimate of the percentage of observers who would vote for B as the more distorted, if a very large number took part in the experiment".

(d) Reliability of the Data

Taking into account all the data (750 observations) obtained in the piano-programme tests, the variance of any observer's rating was found to be in the region of $(0{\cdot}9 \text{ grade units})^2$.

Taking account of this variance, and applying the usual methods, it is quite easy to show that the small variations in the intercept on the Y-axis of the first five diagrams in Fig. 9 are not significant in the statistical sense; that is to say they can be treated as identical.

No satisfactory way was found of testing statistically the validity of the S_{BM} measurements (Fig. 2). The variance of the scores was approximately 1·0 grade unit squared, five grades being involved. 'A'+, 'A'−, 'Equal', 'B'−, 'B'+. Pending a solution, limit-bars were placed above and below each of the plotted points as shown in Fig. 2; they do not define the reliability of the scores but serve as a warning when fitting a line to the data that the position of some points on the diagrams is less certain than others.

In a typical test run, six observers make two observations at any given value of S_B, a single error made by one observer while making these observations, in placing his grading at the adjacent instead of the correct category, would have altered the 'weighted score' by unity. Since in the course of the test the six observers could have registered a maximum possible weighted score (for 'B'+) of 48, this single error of grading would have changed the percentage score by roughly 2%. To indicate this, the bars would lie at $\pm$ 2% around the average percentage score p; this has been done in Fig. 2.

Notice that for scores of very low validity such as p is 98% or 2%, one of the limit bars would lie off the paper at $\pm \infty$.

APPENDIX 4

Characteristics of the Spectrum of a Clipped Sine-Wave

(a) The Spectrum of a Clipped Sine Wave

Fig. 14 (a) is in terms of the quantity d/T, which is defined in the inset 14 (b) in the same figure, and of n^2p_n the amplitudes of the various harmonics after they have each been weighted by the quantity n^2. The sign of the spectral terms may be negative; this indicates only a change of phase. In summing a series of weighted terms in order to derive C_t no account is taken of sign, but in computing C_t for doubly-clipped waves, sign is important as explained below.

The weighted spectrum of a sine-wave clipped by an ideal diode is expressed by

$$n^2p_n = n\,\frac{d}{T}\left[\frac{\sin\,(n-1)\,\frac{d}{T}\,\pi}{(n-1)\,\frac{d}{T}\,\pi} - \frac{\sin\,(n+1)\,\frac{d}{T}\,\pi}{(n+1)\,\frac{d}{T}\,\pi}\right] \times 100 \text{ (per cent)} \quad .. \quad .. \quad (2)$$

where n = harmonic number;

p_n = percentage amplitude ratio of the nth harmonic to the amplitude of originating sine wave (i.e., before clipping).

The value of C_0 is found by summing all terms up to the tenth without consideration of sign; to obtain C_t, n^2t is first subtracted from each term.

(b) Spectra of Waves Clipped Twice

The plotted curves apply to a wave clipped once only as indicated in Fig. 14 (b). For a wave clipped twice (above and below the zero line) two separate tabulations are made as in Table 2. For example, if a wave which had the upper half of the wave clipped at $d/T = 0{\cdot}33$, were clipped a second time (below the line) with $d/T = 0{\cdot}2$, we should obtain the figures tabulated.

The last line (C) of the Table is the spectrum of the doubly-clipped wave and is the sum of (A) and (B)′, the spectra of one wave clipped at $d/T = 0{\cdot}33$ and of the other clipped at $d/T = 0{\cdot}20$, the latter spectrum having signs of the *even terms reversed*.

When the clipping is symmetrical, lines A and B are identical except for signs; so line (C) has no even terms, and the odd terms are doubled.

(c) The Relationship between d/T and q, the Degree of Clipping

A simple geometric relationship exists between the quantity d/T and the degree of clipping q illustrated in Fig. 14 (b):

Since d represents an angle, say α, it follows that h in the figure is given by $h = [1-\cos(\alpha/2)]\,A$, where A is the amplitude of the wave. $q = h/A$, and is the fraction clipped off the wave. The slope of the difference signal is $2h/d$.

APPENDIX 5

On the Increased Unpleasantness of a Distorted Sound when Dispersed

The experiments to be described were made at the stage at which there was ample evidence that the unpleasantness varied in sympathy with changes in the slope of the difference signal. To bring this evidence into line with the facts of the situation, it was necessary to assume that the picture represented by the voltage/time curve measured at the terminals of the distorting diode was transmitted to the observer as some specific pattern of spectral distortion terms, and that the recognition of the pattern was achieved by a process which was immune to the quite gross phase-shifts which these terms must have suffered in generation and acoustical transmission.

The following experiments showed that the immunity existed if the phase-shift was small, but that if it was large the unpleasantness of the sounds *increased* very considerably.

To obtain the data illustrated in Fig. 23 (a) the A and B signals were heavily distorted, the B signal was in addition passed through an all-pass passive network with a non-linear phase response shaped to delay the fundamental term of the spectrum by an adjustable amount (from about 1 to 30 msec). The higher frequency terms (which represent the distortion per se) were unaffected and therefore arrived earlier than the fundamental. It should be mentioned at once that when this situation was reversed, by playing a tape-recording of the signals backwards, the reversal of time sequence made no difference whatever to the listeners' reports. Again, the introduction of the delay-network *made no audible change at all* in the sound heard when the distorted signal was fed through it as a *steady tone* instead of in pulses.

The phenomenon was not fully explored once the general nature of the effect was established; most of the tests were made at 465 or 565 c/s. After a few experiments at lower and higher frequencies, a very rough estimate was made [Fig. 23 (b)] of the way in which the maxima of the curves of Fig. 23 (a) varied with frequency. The close similarity between this curve and Fig. 5 suggests that the same neural/mental process is operative. It is thought that careful repetitions of such experiments could be valuable in the investigation of the theory of hearing.

The degree of non-linearity (the fraction q) had to be held constant throughout the series so far as the undelayed A signal was concerned, and q had to be large (near 0·5) in order to make the unpleasantness significant at 250 c/s; it followed that at six times this frequency

TABLE 2
Computation of Spectrum of Doubly-Clipped Wave

d/T	Values of n	2	3	4	5	6	7	8	9	10
0·33	Values of n^2p_n	53	61	26	−24	−56	−24	23	55	30 (A)
0·20		17	32	40	39	28	10	−13	−31	−38 (B)
Reverse Evens		−17	32	−40	39	−28	10	13	−31	38 (B)′
Add (A) and (B)		36	93	−14	15	−84	−14	36	24	68 (C)

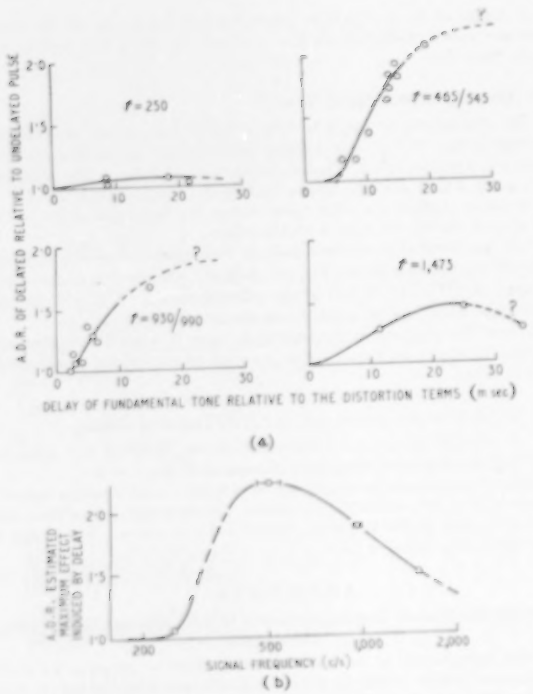

A.D.R. OF DELAYED RELATIVE TO UNDELAYED PULSE
f = 250
f = 485/545
f = 930/990
f = 1,475
DELAY OF FUNDAMENTAL TONE RELATIVE TO THE DISTORTION TERMS (m sec)
(a)
A.D.R. ESTIMATED MAXIMUM EFFECT PRODUCED BY DELAY
SIGNAL FREQUENCY (c/s)
(b)

Fig. 23. *Increase of unpleasantness arising from delay of fundamental component*

(1,500 c/s) the distortion was extremely unpleasant and critical judgement was difficult. The unpleasantness was mitigated by diluting this A signal ($S_A = 6$) throughout the series of tests.

The existence of this delay phenomenon justifies the addition of the terms of the weighted distortion spectrum *without consideration of sign* in arriving at the distortion criterion C_D for a change of sign can be equated to 180° phase shift and this is but a small fraction of the shift shown here to be unrecognizable subjectively (two complete periods of the wave).

Again, the deviation-rate hypothesis receives support if the following explanation is accepted: We suppose that when there is no delay, the 'gestalt' (the overall mental picture of the acoustic situation which is built up in the listener's mind) contains an imagined sine-wave signal which is identical with the wave from which the distortion was generated; such a proposition is basic to the hypothesis. We suppose further, that the gestalt is completely modified by the separation-in-time of the distortion and fundamental terms of the spectrum which occurs during the delay-experiment, and further that, as a result of the breakdown, the fundamental term recognized by the listener reverts to its 'real' value; i.e., what would be measured by a wave-analyser.

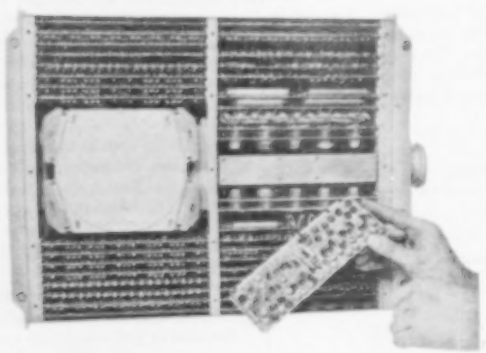
This real term is, of course, always smaller than the previously imagined one, especially is this the case when non-linearity distortion has been made severe, as here. It follows that the listener's appreciation of the distortion alters: he finds that it becomes more unpleasant, because the distortion terms are now contrasted with a smaller fundamental term. Thus, it is possible to explain the increase of a.d.r. which is associated with an increase of delay in Fig. 23.

The phenomenon just dealt with can be strikingly and simply demonstrated *in reverse*; in so far as delay, or separation in time, of the distortion terms relative to the fundamental causes unpleasantness to increase so, also, the *removal* of any already existing delay or time-separation should *reduce* the unpleasantness. This can be demonstrated by listening to distorted sounds through high-grade headphones instead of a loudspeaker; the latter has to radiate into a room which necessarily imposes frequency-selective delay upon the sounds heard.

An elementary experiment was done on these lines, using the tape recording of the piano programme employed earlier ("The Audience Reaction to Distorted Programme Material"; Part 1). It was found that the gradings assigned by the observer dropped in the ratio of 3 to 5 when the headphones were used in place of the loudspeaker—a ratio which was so large as to demand artificial-ear tests to prove that the headphones had, in fact, an unimpeachable frequency-response.

ELLIOTT BROS. TO MARKET U.S. COMPUTER

The Autonetics Division of North American Aviation Inc., and Elliott Brothers (London) Ltd., have entered into an agreement which allows Elliott to manufacture and market the Verdan digital computer in the U.K. and the Commonwealth.



The Verdan machine is a serial general-purpose digital computer combined with a serial digital differential analyser and is designed for use in guidance and control systems of American aircraft, missiles and submarines. These two sections interact simultaneously to solve a variety of in-flight problems and generate appropriate command signals in 'real-time' for use with other sub-systems.

A special feature of the design of this machine is that its power may be interrupted during a computation without loss of stored information; this is accomplished by a design technique wherein the computer automatically marks its place in a computation whenever a power failure is detected and retains stored information until power is restored and the computation is resumed.

Internal storage is on a magnetic disc (operating at 6,000 r.p.m.) with a capacity of 1,664 words in 13 channels; the words comprise 26 bits, including sign, spare and synchronizing bits.

The illustration shows the plug-in module construction used and indicates the size of the machine.

NEW AGENCIES FOR CONTINENTAL DISTRIBUTORS

Continental Distributors Ltd., have recently announced two new agency agreements with overseas companies. The first is with Alderson Research Laboratories, Long Island City, New York, U.S.A. Alderson Research are manufacturers of a range of radiological test phantoms and anthropomorphic test dummies. (Test phantoms are plastic dummies with a skeleton, cells and tissue solution, which may be used to simulate the human body for certain radio-activity absorption tests. Anthropomorphic dummies are mechanical models of the 'average man', which may be used to test the physical reaction of the human body to mechanical stresses).

The second agreement announced is with Sanser, Elettronica Per Fisica Nucleare, 225 Viale G. Suzzani, Milan, Italy, for the sales and servicing in the U.K. of their automatic bi-dimensional radiochromatogram scanner.

Matrices: Pitfalls and Further Examples

In the January 'Tool', a number of circuits were discussed where matrices greatly assisted the task of specifying broad relations between those circuits (or major portions of them) and the external world; it was seldom necessary for us to concern ourselves with internal currents and voltages in the circuit. In general, however, two basic assumptions were made in this article, implied by the fact that all the circuits there discussed could be classed as three-terminal networks. These assumptions were:

(a) If the general network is specified in terms of the notation of Fig. 1, the voltage at the terminal 2′ is the same as that at the terminal 1′.

(b) The current I_1 entering the network at the terminal 1 is equal and opposite to the current leaving the network at the terminal 1′. [Assumption (b) implies that the current I_2 leaving the network at terminal 2 is also equal and opposite to that entering the network at the terminal 2′.]

These assumptions did not enter explicitly into any calculations, and once we are satisfied that they apply to any network with which we are concerned, they may be forgotten while matrix techniques are applied. Matrix techniques discussed in various 'Mathematical Tools' are not, in fact, applicable to the general four-terminal network but, in practice, this limitation is

Fig. 1. *A general network*

more apparent than real, and the vast majority of networks in use are suitable for the application of matrix techniques.

In communication technique, a single line is commonly protected from interference by a concentric screen. This is an unbalanced system because the outer is earthed. An alternative way of reducing interference is by means of a symmetrically-balanced pair. In this case it is expected that stray elements occurring on one line will cancel out equal stray elements occurring on the other. Currents going along both lines alike can be called 'push-push'; those which are equal and opposite along the two lines can correspondingly be called 'push-pull'. A balanced structure uses only the 'push pull' component, and is usually associated with transformers or other terminal equipment designed to ignore the unwanted 'push-push' component and accept the 'push-pull'. Occasionally such structures are used in both 'push-pull' and 'push-push' modes, but this is not in general of great interest and will not be considered further here. The structures which are of interest are either balanced in the 'push-pull' mode, or unbalanced, like the concentric cable, with a common connection or earth. Either of these has the right

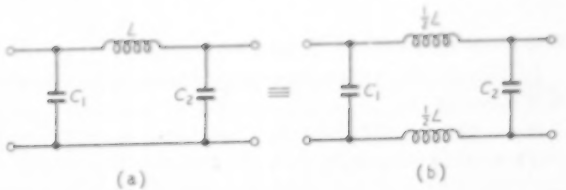


Fig. 2. *(a) A simple three-terminal network; (b) balanced equivalent of (a)*

number of degrees of freedom for the application of the 2×2-matrix techniques we have considered.

Balanced circuits need not be considered separately from three-terminal networks, for (to take a simple example) the three-terminal unbalanced circuit of Fig. 2 (a) is equivalent to the balanced circuit of Fig. 2 (b), and thus a mere minor reorganization is sufficient to permit almost all useful work to be done in terms of three-terminal networks. Certain structures (such as lattices) appear not to be amenable to this treatment, but the use of such elements implies the presence of terminal transformers to look at the push-pull component only. The corresponding three-terminal network can be obtained in various ways by means of transformers and mutual coupling; for example, as in Fig. 3. For purposes of thought and understanding, it is desirable to replace a lattice by an equivalent unbalanced structure, as, for example, in Fig. 4. The lattice (and

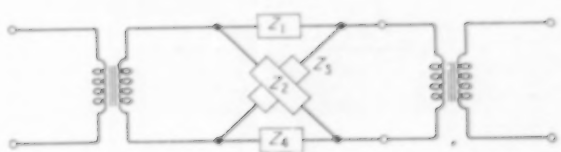


Fig. 3. *Lattice network with associated transformers to select push-pull component*

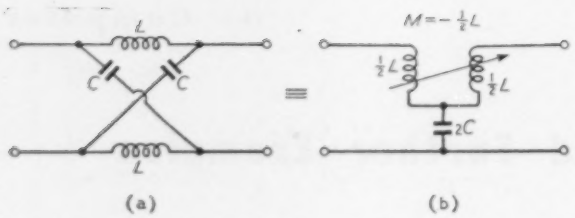


Fig. 4. *(a) Lattice network; (b) unbalanced (three-terminal) equivalent*

bridged-T) structures come into the category of two-path networks, because they can be realized by connecting two ladder structures in parallel-parallel. The essential quality of a lattice is that it should be equivalent to a straight-through connection at one frequency and a pole change at another frequency.

Thus here we concern ourselves only with the application of matrix techniques to networks which can be regarded as in equivalent three-terminal form. The remainder of this article will be devoted to matrices associated with non-ideal transformers, valves and transistors.

Matrices Applied to Transformers

In Fig. 5, the easiest matrix to obtain direct is the Z-matrix, since

$$\left.\begin{aligned} V_1 &= pL_1I_1 - pMI_2 \\ V_2 &= pMI_1 - pL_2I_2 \end{aligned}\right\} \quad .. \quad .. \quad .. \quad (1)$$

or

$$Z = \begin{pmatrix} pL_1 & -pM \\ pM & -pL_2 \end{pmatrix} \quad .. \quad .. \quad .. \quad .. \quad (2)$$

The conversion formulae given in the April 1958 'Tool' give us immediately the corresponding A-matrix

$$A = \begin{pmatrix} L_1/M & p(L_1L_2 - M^2)/M \\ 1/pM & L_2/M \end{pmatrix} \quad .. \quad .. \quad (3)$$

It will save space to write $x = L_1L_2 - M^2$; we should note that if x is zero (so that the coupling in Fig. 5 is perfect) then for sufficiently high frequencies (so that the 'shunt-ness' element $1/pM$ can be neglected) A in Equ. (3) is the matrix of an ideal transformer of ratio $L_1/M = (L_1/L)^{1/2}$ in this case and the elements of the Z-matrix become infinite. In Fig. 6 (a) the same basic Equs. (1) are the easiest to derive directly, and thus the circuits of Figs. 5 and 6 are equivalent at all frequencies. In the case of Fig. 6 (b) however, we have

$$A = \begin{pmatrix} 1 & 0 \\ \dfrac{L_2-M}{px} & 1 \end{pmatrix} \begin{pmatrix} 1 & px/M \\ 0 & 1 \end{pmatrix} \begin{pmatrix} 1 & 0 \\ \dfrac{L_1-M}{px} & 1 \end{pmatrix} \quad (4)$$

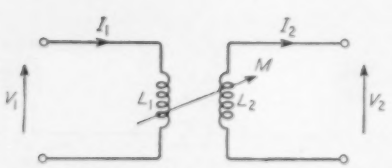


Fig. 5. *A coupled circuit*

and this leads directly to Equ. (3). Thus, the effect of a coupled circuit can be allowed for by means of the circuits of Figs. 6 (a) and 6 (b) which are not effectively different from circuits without coupling. Further, there is no reason why, when a coupled circuit is required, conceived initially as being of the form of Fig. 5, it should not be realized in the form of Fig. 6 (a) or 6 (b), and vice versa.

In Fig. 7, we have likewise a circuit that can be made equivalent to that of Fig. 5 at any one frequency for fixed C_2, and over a whole range of frequencies if C_2 can be varied. For in Fig. 7

$$\begin{pmatrix} V_1 \\ I_1 \end{pmatrix} = \begin{pmatrix} 1 & 0 \\ 1/pl_1 & 1 \end{pmatrix} \begin{pmatrix} 1 & 1/pC_2 \\ 0 & 1 \end{pmatrix} \begin{pmatrix} 1 & 0 \\ 1/pl_3 & 1 \end{pmatrix} \begin{pmatrix} V_2 \\ I_2 \end{pmatrix} \quad (5)$$

so that

$$A = \begin{bmatrix} 1 + l_1/(l_3y) & pl_1/y \\ (1/pl_1)+(1/pl_3)+(1/pl_3y) & 1 + (1/y) \end{bmatrix} \quad (6)$$

where

$$y = p^2l_1C_2 \quad .. \quad .. \quad .. \quad .. \quad .. \quad (7)$$

and this is the same as in Equ. (3) if

$$l_1 = \frac{x}{L_2-M}\ ; \quad l_3 = \frac{x}{L_1-M}\ ; \quad y = \frac{M}{L_2-M} \quad .. \quad (8)$$

but of course y varies with frequency, from Equ. (7),

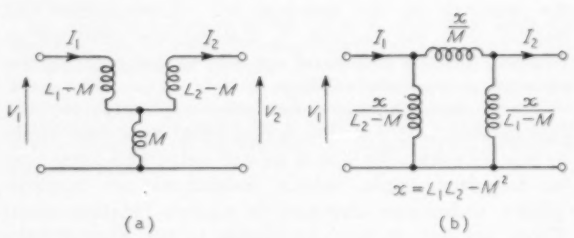


Fig. 6. *(a) and (b) Circuits equivalent to the coupled circuit of Fig. 5, but without mutual coupling*

unless C_2 can be varied appropriately. Thus, if we think of coupled circuits as having the A-matrix of Equ. (3) and as realizable in the form of Figs. 6 (a), 6 (b) or 7, the sting is taken out of them; they can be handled just in the same way as circuits without coupling.

Matrices Applied to Valves and Transistors

Given a triode with anode resistance r_a and mutual conductance μ, the essential facts of the situation are expressed by means of the relations

$$\begin{aligned} -\mu V_1 &= V_2 + r_a I_2 \\ I_1 &= 0 \end{aligned}$$

in the notation of Fig. 8. Translated into matrix language, this means that a valve has an A-matrix just like any other network, active or passive, namely

$$A_v = \begin{bmatrix} -1/\mu & -r_a/\mu \\ 0 & 0 \end{bmatrix} \quad .. \quad .. \quad .. \quad .. \quad (9)$$

The only peculiarity about A_v in Equ. (9) is that $|A_v| = 0$, and there is a very definite physical significance to this in that a triode arranged as in Fig. 8 is a one-way device: I_1 is always zero. But once Equ. (9) is estab-

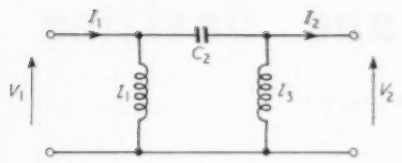


Fig. 7. *A circuit which can be made equivalent to that of Fig. 5 (at any one frequency for fixed C_2)*

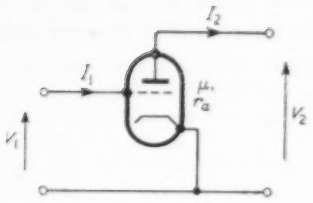


Fig. 8. *An idealized triode*

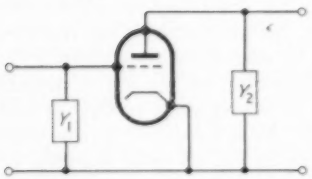


Fig. 9. *Shunt admittances connected to the idealized triode of Fig. 8*

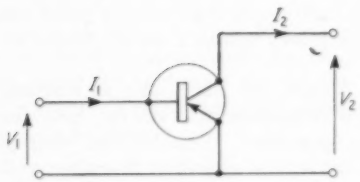


Fig. 10. *Common-emitter transistor corresponding to the idealized triode of Fig. 8*

lished, circuits involving triodes can be understood in terms of matrices just like any other circuits; for instance, if the triode is preceded by an admittance Y_1 in shunt as in Fig. 9, and followed by a similar admittance Y_2, we have

$$A = \begin{bmatrix} 1 & 0 \\ Y_1 & 1 \end{bmatrix} \begin{bmatrix} A_v \end{bmatrix} \begin{bmatrix} 1 & 0 \\ Y_2 & 1 \end{bmatrix} \quad .. \quad .. \quad .. \quad (10)$$

so that, in Fig. 9,

$$\begin{pmatrix} V_1 \\ I_1 \end{pmatrix} = \begin{pmatrix} -(1 + r_a Y_2)/\mu & -r_a/\mu \\ -Y_1(1 + r_a Y_2)/\mu & -Y_1 r_a/\mu \end{pmatrix} \begin{pmatrix} V_2 \\ I_2 \end{pmatrix} \quad (11)$$

and it is easily verified that the determinant of the A-matrix on the right-hand side of Equ. (11) is zero, although all the elements are different from zero. This we should naturally expect, since the one-way character of the idealized triode of Fig. 8 is not affected by the preceding passive elements introduced in Fig. 9.

Thus once Equ. (9) is established, we are at liberty to forget that A_v is associated with a valve and to treat it like any other A-matrix. The only thing we cannot do is to use A_v^{-1}, the inverse of A_v, but this is an eminently sensible prohibition, since using a one-way device the wrong way is even more crazy than driving a car in the wrong direction up a one-way street.

Transistors, like valves, can be associated with A-matrices having determinant zero, and there is a dual relation between these matrices for valves and transistors, as pointed out by Mr. Wilson in a recent letter (February issue, p. 69). The common-emitter transistor (idealized as in Fig. 10, that is, with emitter and base resistances neglected) has the A-matrix which is dual to that of Equ. (8), namely

$$A_T = \begin{bmatrix} 0 & 0 \\ -g_c'/\alpha_a' & -1/\alpha' \end{bmatrix} \quad .. \quad .. \quad .. \quad (12)$$

where

$$\left.\begin{array}{l} g_c \text{ is the collector conductance} \\ \alpha \text{ is the current gain} \\ g_c' = g_c/(1-\alpha) \\ \alpha' = \alpha/(1-\alpha) \end{array}\right\} \quad .. \quad .. \quad (13)$$

Thus matrices can be used to deal with almost any circuit situation which can be expressed in terms of three-terminal networks. Kirchhoff's laws can be used to obtain correct results if no error is made in the rather heavy algebra usually involved, but it is very difficult to see the significance of results obtained in this way without much hard labour. Matrix technique automatically introduces some order and grouping into the analysis, and greatly reduces the possibility of making silly mistakes because it is difficult in a forest of symbols to find one's way and be sure what one is trying to do. A result which is complicated even when matrices are used is hardly likely to be simple otherwise.

NEW TELEVISION CAMERA

A small cylindrical closed-circuit television camera is now being manufactured by E.M.I. Measuring just over 9 inches from lens-mount to back-plate and 3 in. diameter, it uses a 1-in. vidicon tube.

To achieve the small size, as few components as possible have been placed in the camera and, as a result, no ventilation holes which would allow the ingress of dust are necessary.

Theory of Waveguides and Cavities

3: PERTURBATION THEORY AND ITS APPLICATIONS

By R. A. Waldron, M.A. (Cantab.), A. Inst. P., A.M. Brit. I.R.E.*

SUMMARY. *The bases, applications, and limitations of perturbation theory are discussed in relation to waveguides and cavities. Two examples are treated—the use of a cavity to measure the dielectric constant of a solid material, and the attenuation in a waveguide due to the dielectric loss. In the latter case, the result is compared with that obtained in the second article of this series for the same problem, using an exact method.*

In the first two articles of this series[1], we have seen how the choice of a model influences the kinds of results that are obtainable when treating a physical problem. By choosing a simple model, a limited amount of information is obtained, while for a more sophisticated model more information can be obtained. For example, if the metal wall of a waveguide is taken to be a perfect conductor, the characteristic equation can be solved for a phase constant, β, which gives no information about the effect of finite conductivity of the metal parts of the guide. But if a model is taken which takes the metal to be of finite conductivity, this further information is implicit in the characteristic equation.

For any practical waveguide, it is necessary to choose a model that departs in some respect from the physically existing waveguide, and the characteristic equation can therefore not give all the information that is likely to be required. With certain limitations, the extra information can usually be obtained by the use of perturbation theory. This can be used to calculate the effect on the phase constant of a waveguide, or on the frequency of a cavity, of a departure from the mathematical model considered. It is essential that such a departure shall be small, but this must be the case if the model has been well chosen. If the model is not well chosen, the phase constant for the model will not approximate to that of the actual guide, or the model frequency to the actual cavity frequency, and the result is not of much value.

The essence of the perturbation calculation is to determine a small change in the phase constant of a waveguide, or in the resonance frequency of a cavity, due to a small departure of the physical system from the model used. Thus the perturbation method depends on previously having obtained a solution, for the mode in question, of the characteristic equation. This is one limitation of the method; it is not always possible to obtain, from the characteristic equation for the model, the eigenvalue for a certain mode of the actual system—the phase constant of a guide or the frequency of a cavity. An example of this was given in the second article of this series; the HSP model, for the waveguide which consists of a tube of metal of circular cross-section containing air, does not enable the E_{p0} modes to be treated. The results given by the more general treatment, taking into account the finite conductivity of the metal and the loss tangent of the air, do enable the E_{p0} modes to be investigated. It is not possible to obtain any information about these modes by applying perturbation theory to the results for the HSP model. Putting it another way, before perturbation theory can be used, there has to be something to perturb.

Another limitation on the use of perturbation theory in the case of waveguides is that it cannot be used near cut-off. As a property of the waveguide varies from its value at cut-off (e.g. working frequency, one of the dimensions, dielectric constant of one of the media, etc.), the phase constant usually changes rapidly at first, so that quite a small departure, physically, of the physical system from the model can cause a considerable change of the phase constant. Perturbation theory, however, is only applicable when a small change in the physical conditions causes a small change in the eigenvalue.

In this article, we shall take as starting point the perturbation formulae and discuss how they may be applied in practice. One of the examples treated exactly in the second article of this series will be treated by the perturbation method, and the results compared with those given previously.

The Perturbation Formulae

It is convenient to deal first with the perturbation of cavities, and then to develop this theory to cover the waveguide case. A perturbation consists in changing the material in some part of the cavity. The change may be to replace, in a small volume, the material present in the unperturbed condition by some other material differing considerably in properties. This is the case when a cavity method is used for the measurement of the electric and magnetic properties of small solid samples, and it can also be used to treat small changes of shape in the cavity by regarding the change of shape as consisting in the replacement of air by metal or metal by air. Alternatively, the change may consist in replacing, over a large volume (perhaps the whole volume of the cavity), the material present in the

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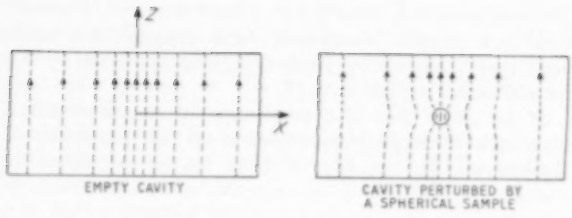

Fig. 1. *Electric field in a rectangular cavity excited in an E_{110} mode; section through the centre of the cavity in the $x - z$ plane*

unperturbed condition by some other material differing only slightly in properties. This is the case when a cavity method is used to measure the electric and magnetic properties of a gas.

Let us suppose that, in the unperturbed condition, the electric and magnetic fields in the cavity are $\mathbf{E}_0$ and $\mathbf{H}_0$ in some mode on which we fix our attention. $\mathbf{E}_0$ and $\mathbf{H}_0$ are supposed known. We also write $\mathbf{D}_0$ and $\mathbf{B}_0$ for the inductions, and let ω_0 be the resonance frequency of the cavity. In the perturbed condition, let the fields become $\mathbf{E}_0 + \mathbf{E}_1$ and $\mathbf{H}_0 + \mathbf{H}_1$, and let the resonance frequency become $\omega_0 + \delta\omega$. $\mathbf{E}_1$ and $\mathbf{H}_1$ are fields that must be added to the unperturbed fields to give the fields in the perturbed cavity. There may be a change in scale factor between the fields in the perturbed and unperturbed conditions; $\mathbf{E}_1$ and $\mathbf{H}_1$ are not intended to take account of this. $\mathbf{E}_0$ and $\mathbf{H}_0$, in the perturbed condition, represent all the energy in the mode in question, while $\mathbf{E}_1$ and $\mathbf{H}_1$ can, in principle, be represented as a sum of all modes except the one obtaining in the unperturbed cavity. We also write $\mathbf{D}_0 + \mathbf{D}_1$ and $\mathbf{B}_0 + \mathbf{B}_1$ for the inductions in the perturbed cavity.

For the perturbation theory to be valid, it is necessary that over nearly all of the volume of the cavity $\mathbf{E}_1$, $\mathbf{H}_1$, $\mathbf{D}_1$, and $\mathbf{B}_1$ are small compared with $\mathbf{E}_0$, $\mathbf{H}_0$, $\mathbf{D}_0$, and $\mathbf{B}_0$, respectively, although this may not be the case in a very small volume. The perturbation formulae for the cavity may then be stated:

Case 1: Large Change in Properties over Small Volume

$$\frac{\delta\omega}{\omega_0} = \frac{\iiint_{V_1} \{(\mathbf{E}_1.\mathbf{D}_0 - \mathbf{E}_0.\mathbf{D}_1) - (\mathbf{H}_1.\mathbf{B}_0 - \mathbf{H}_0.\mathbf{B}_1)\}\, dV}{\iiint_{V_0} (\mathbf{E}_0.\mathbf{D}_0 - \mathbf{H}_0.\mathbf{B}_0)\, dV} \quad \ldots \quad \ldots \quad (1)$$

Case 2: Small Change in Properties over Large Volume

$$\frac{\delta\omega}{\omega_0} = \frac{\iiint_{V_0} \{(\mathbf{E}_1.\mathbf{D}_0 - \mathbf{E}_0.\mathbf{D}_1) - (\mathbf{H}_1.\mathbf{B}_0 - \mathbf{H}_0.\mathbf{B}_1)\}\, dV}{\iiint_{V_0} (\mathbf{E}_0.\mathbf{D}_0 - \mathbf{H}_0.\mathbf{B}_0)\, dV} \quad \ldots \quad \ldots \quad (2)$$

A detailed derivation of Equ. (1) has been given[2].

These formulae are first-order approximations which hold to a fairly high degree of accuracy, provided that the conditions under which they are obtained are observed. In case 1, V_1 is the volume of a small specimen, very much less than V_0, the volume of the cavity. $\mathbf{E}_1$, $\mathbf{H}_1$, $\mathbf{D}_1$, and $\mathbf{B}_1$ may depart considerably from zero over the volume V_1, and outside the sample but still in its neighbourhood, as long as they are nearly zero everywhere else. In case 2, $\mathbf{E}_1$, $\mathbf{H}_1$, $\mathbf{D}_1$, and $\mathbf{B}_1$ must be small everywhere in the cavity. If these conditions are observed, then $\delta\omega/\omega_0$ will be small; in practice, only small frequency shifts are observed, and Equs (1) and (2) are then valid.

The evaluation of the denominators of Equs (1) and (2) presents no difficulty; $\mathbf{E}_0$, $\mathbf{D}_0$, $\mathbf{H}_0$ and $\mathbf{B}_0$ are all known, so that the integration can be carried out. For the numerator, it is necessary to perform a subsidiary calculation to determine $\mathbf{E}_1$, $\mathbf{H}_1$, $\mathbf{D}_1$ and $\mathbf{B}_1$, before the integration can be performed. We shall discuss this question in a later section (The Evaluation of $\mathbf{E}_1$, $\mathbf{H}_1$, $\mathbf{D}_1$, $\mathbf{B}_1$). Before doing this, however, we now consider the application of the perturbation formulae to waveguides.

For a waveguide, let the volume V_0 be the volume of a length of guide equal to one guide wavelength, and let the frequency of waves supplied by a generator be ω_0. This generator is supposed to be at infinity, so that the guide can be considered to be working in a single mode. Now let a perturbation be applied, e.g. the guide is filled with a different gas, or a thin rod of dielectric material is placed at its centre, or its shape is slightly deformed. At the same time, let the frequency be changed to $\omega_0 + \delta\omega$ so that at this new frequency the guide wavelength is unchanged. The situation in the volume V_0 is now analogous to that in the cavity—in fact, the guide can be regarded as an infinitely long cavity. The perturbation is of the cross-section only—the changes made are made over the whole length of the guide. Thus in both numerator and denominator of the perturbation formulae, the integration with respect to z gives the same multiplying factor, which can be cancelled. It is therefore only necessary to integrate over the cross-section, and in place of Equs (1) and (2) we have:

Case 1: Large Change in Properties over Small Cross-Sectional Area

$$\frac{\delta\omega}{\omega_0} = \frac{\iint_{S_1} \{(\mathbf{E}_1.\mathbf{D}_0 - \mathbf{E}_0.\mathbf{D}_1) - (\mathbf{H}_1.\mathbf{B}_0 - \mathbf{H}_0.\mathbf{B}_1)\}\, dS}{\iint_{S_0} (\mathbf{E}_0.\mathbf{D}_0 - \mathbf{H}_0.\mathbf{B}_0)\, dS} \quad \ldots \quad \ldots \quad (3)$$

Case 2: Small Change in Properties over Large Cross-Sectional area

$$\frac{\delta\omega}{\omega_0} = \frac{\iint_{S_0} \{(\mathbf{E}_1.\mathbf{D}_0 - \mathbf{E}_0.\mathbf{D}_1) - (\mathbf{H}_1.\mathbf{B}_0 - \mathbf{H}_0.\mathbf{B}_1)\}\, dS}{\iint_{S_0} (\mathbf{E}_0.\mathbf{D}_0 - \mathbf{H}_0.\mathbf{B}_0)\, dS} \quad \ldots \quad \ldots \quad (4)$$

In the case of the waveguide, however, it is usually required to keep the frequency fixed and to calculate the change in phase constant. We thus consider a further change; in the perturbed state, the frequency is returned to ω_0, and correspondingly the phase constant

changes from β_0, its value in both the unperturbed state at frequency ω_0 and in the perturbed state at frequency $\omega_0 + \delta\omega$, to $\beta_0 + \delta\beta$. Then

$$\delta\beta = -\left(\frac{d\beta}{d\omega}\right)_{\omega_0} \delta\omega \quad .. \quad .. \quad .. \quad .. \quad (5)$$

$$\text{i.e., } \frac{\delta\beta}{\beta_0} = -\left(\frac{d\beta}{d\omega}\right)_{\omega_0} \frac{\delta\omega}{\omega_0}\frac{\omega_0}{\beta_0} \quad .. \quad .. \quad .. \quad (6)$$

The minus sign arises because the change in frequency is now $-\delta\omega$. It is sometimes convenient to write

$$\left(\frac{d\beta}{d\omega}\right)_{\omega_0} = \frac{d\beta_0}{d\lambda_0}\left(\frac{d\lambda_0}{d\omega}\right)_{\omega_0} \quad .. \quad .. \quad .. \quad (7)$$

in order to evaluate Equ. (6), λ_0 being the free-space wavelength.

The Evaluation of $\mathbf{E}_1$, $\mathbf{H}_1$, $\mathbf{D}_1$, $\mathbf{B}_1$

The evaluation of these quantities is not always possible, and then the perturbation method breaks down. Even where it is possible, the calculation may be difficult, but the difficulty is offset by the fact that once the calculation is made it can be applied in a large number of cases.

In measuring the properties of small samples of solid material, it is convenient to place the sample at a point in a cavity where either the electric or the magnetic field is zero; the magnetic or electric properties, respectively, can then be examined separately. It happens that when one field is zero, the other is a maximum, and is therefore approximately uniform over a limited region. If the sample is small compared with this region, the field inside it can be calculated as if it were due to a uniform external field. (By external is meant not the field outside the sample, but the field that existed at the position of the sample before the sample was introduced. The word 'external' is justified because this is still the value of the field at large distances —i.e., it is the field external to the *neighbourhood* of the sample rather than to the sample itself.) The calculation of the internal field can be performed only if the specimen is in the form of an ellipsoid; special cases of the general ellipsoid that are commonly used are the sphere and the rod of circular cross-section. Solutions have been given for the electric and magnetic fields in dielectric and ferrite specimens of a variety of shapes, in terms of the external field[3]. One other case that is tractable is the rod of any cross-section, if the (uniform) microwave field is in the direction of its axis, when the field in the specimen is equal to the external field. An example of the calculation for a small solid sample is given in the following section.

The case of a small change in properties over a large volume is easier to treat. For example, it may be required to calculate the attenuation coefficient of a waveguide consisting of a metal tube containing air, due to the dielectric loss of the air. The perturbation is the change, over the whole of the volume of the waveguide (if the metal is regarded as a perfect conductor), of the dielectric constant from 1 to $1 - j\epsilon''$. $\mathbf{E}$, $\mathbf{H}$, and $\mathbf{B}$, are unchanged, so that $\mathbf{E}_1$, $\mathbf{H}_1$, and $\mathbf{B}_1$ are zero. $\mathbf{D}_1$ is easily calculated, since

$$\mathbf{D}_0 + \mathbf{D}_1 = \epsilon_0\epsilon\mathbf{E}_0 = \epsilon_0(1 - j\epsilon'')\mathbf{E}_0$$
$$\therefore \mathbf{D}_1 = \epsilon_0(1 - j\epsilon'')\mathbf{E}_0 - \epsilon_0\mathbf{E}_0 = -j\epsilon''\epsilon_0\mathbf{E}_0$$

We shall treat this example in a later section (Dielectric Loss in a Simple Waveguide), and compare the result with that obtained by a more general approach in the second article of this series.

It is clear that the possibility of a perturbation approach to a problem depends on the possibility of determining $\mathbf{E}_1$ and $\mathbf{H}_1$; if these can be calculated, so can $\mathbf{D}_1$ and $\mathbf{B}_1$. The kinds of problem that are tractable are therefore the calculation of the frequency shift of a cavity due to the placing of a small ellipsoidal specimen in a region of uniform field; the calculation of the frequency shift of a cavity due to the introduction of a gas which fills the cavity; the calculation of the finite Q of a cavity due to the finite conductivity of the walls; the calculation of the frequency shift of a cavity due to a small deformation of its surface, if the deformation is of a suitable shape; and the calculation of the change of phase constant of a waveguide due to the same kinds of changes. In the waveguide case, the change must be uniform over the whole length of the guide.

We may note here that the well-known method of treating losses in waveguides due to the finite conductivity of the walls (see, e.g., references 4, 5 and 6) is an example of the application of perturbation theory, although it is not usually stated to be such.

Frequency Shift of a Cavity Due to a Small Solid Specimen

As an example, consider a sphere of dielectric material of relative permittivity ϵ, placed in a cavity in such a way that the external microwave electric field is uniform and transverse in the neighbourhood of the sphere. One way of achieving this situation is to use a rectangular cavity in an E_{mn0} mode, with m and n odd and the sample at the centre of the cavity. In the E_{mn} modes, the field components are

$$\left.\begin{aligned}
E_z &= \sin\frac{m\pi x}{a}\sin\frac{n\pi y}{b}\cos\frac{\nu\pi z}{c}\\
E_x &= \frac{-1}{k^2}\frac{m\pi}{a}\frac{\nu\pi}{c}\cos\frac{m\pi x}{a}\sin\frac{n\pi y}{b}\sin\frac{\nu\pi z}{c}\\
E_y &= \frac{-1}{k^2}\frac{n\pi}{b}\frac{\nu\pi}{c}\sin\frac{m\pi x}{a}\cos\frac{n\pi y}{b}\sin\frac{\nu\pi z}{c}\\
H_z &= 0\\
Hx &= \frac{j\omega\epsilon_0}{k^2}\frac{n\pi}{b}\sin\frac{m\pi x}{a}\cos\frac{n\pi y}{b}\cos\frac{\nu\pi z}{c}\\
H_y &= \frac{-j\omega\epsilon_0}{k^2}\frac{m\pi}{a}\cos\frac{m\pi x}{a}\sin\frac{n\pi y}{b}\cos\frac{\nu\pi z}{c}
\end{aligned}\right\} \quad .. \quad (8)$$

where a, b, and c are the dimensions of the cavity in the x, y, and z directions respectively. For $\nu = 0$, E_x and E_y vanish identically. For odd m and n, $\cos(m\pi x/a)$ and $\cos(n\pi y/b)$ vanish at the centre of the cavity, where $x = a/2$, $y = b/2$. At the centre, therefore, the only field component is E_z, which is a maximum and approximately uniform.

If a sphere of dielectric material is immersed in an electric field which in the absence of the sphere is uniform and equal to $\mathbf{E}$, the field $\mathbf{E}'$ in the sphere is in the same direction as $\mathbf{E}$, and of magnitude

$$E' = \frac{3E}{\epsilon + 2} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (9)$$

This formula holds for microwave fields as long as the sphere is sufficiently small for the field to be sensibly uniform over a region large compared with the sphere. From this, $\mathbf{E}_1$ and $\mathbf{D}_1$ can be calculated and used in Equation (1).

$\mathbf{E}$ is the value of $\mathbf{E}_0$ at the position of the sphere, i.e., in the volume V_1. $\mathbf{E}'$ is the value of $\mathbf{E}_0 + \mathbf{E}_1$ in V_1. Hence

$$\mathbf{E}_1 = \mathbf{E}' - \mathbf{E}_0 = \mathbf{E}_0 \left\{ \frac{3}{\epsilon + 2} - 1 \right\} = -\mathbf{E}_0 \left\{ \frac{\epsilon - 1}{\epsilon + 2} \right\} \quad .. \quad .. \quad (10)$$

For $\mathbf{D}_1$, we have

$$\mathbf{D}_0 + \mathbf{D}_1 = \epsilon\epsilon_0(\mathbf{E}_0 + \mathbf{E}_1) = \epsilon\epsilon_0 \mathbf{E}_0 \frac{3}{\epsilon + 2}$$

so that

$$\mathbf{D}_1 = \epsilon\epsilon_0 \mathbf{E}_0 \frac{3}{\epsilon + 2} - \epsilon_0 \mathbf{E}_0 = \epsilon_0 \mathbf{E}_0 \, 2 \left\{ \frac{\epsilon - 1}{\epsilon + 2} \right\} \quad .. \quad (11)$$

In the numerator of Equ. (1), the second bracket vanishes because $\mathbf{H}_0$, and therefore $\mathbf{B}_0$, are zero at the

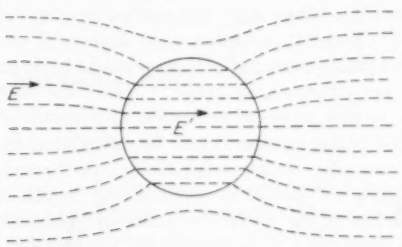

Fig. 2. *Electric field in the neighbourhood of a dielectric sphere, the field being uniform and equal to E in the absence of the sphere. The field in the sphere is given by* $E'(\epsilon + 2) = 3E$, ϵ *being the dielectric constant of the sphere*

position of the sample. From Equs (10) and (11) the first bracket is

$$\mathbf{E}_1.\mathbf{D}_0 - \mathbf{E}_0.\mathbf{D}_1 = \epsilon_0 E_0^2 \left\{ \frac{-(\epsilon - 1)}{(\epsilon + 2)} - 2 \frac{(\epsilon - 1)}{(\epsilon + 2)} \right\} = -3\epsilon_0 E_0^2 \frac{(\epsilon - 1)}{(\epsilon + 2)}$$

Substituting this into Equ. (1), we obtain

$$\frac{\delta\omega}{\omega_0} = \frac{-3(\epsilon - 1)}{(\epsilon + 2)} \frac{\iiint_{V_1} E_0^2 \, dV}{\iiint_{V_0} (\mathbf{E}_0.\mathbf{D}_0 - \mathbf{H}_0.\mathbf{B}_0) \, dV} \quad (12)$$

On performing the integrations, a constant is obtained which depends on the geometry of the cavity and the mode of oscillation. Thus we have

$$\frac{\delta\omega}{\omega_0} = -A \left\{ \frac{\epsilon - 1}{\epsilon + 2} \right\} \quad .. \quad .. \quad .. \quad .. \quad (13)$$

The negative sign indicates that $\delta\omega$ is negative, i.e., on introducing the sphere the frequency is lowered. This is as we should expect, since to increase the dielectric constant of a medium lowers the wavelength, so that relative to the wavelength the space occupied becomes larger.

From Equ. (13) the dielectric constant can be calculated if $\delta\omega/\omega_0$ is measured. Dielectric constants of practical materials are complex, and by regarding the frequency also as complex, Equ. (13) can be applied to the measurement of both real and imaginary parts of ϵ. Let Q_0 be the Q of the cavity in the unperturbed condition, and let us define its complex resonance frequency, Ω_0, by

$$\Omega_0 = \omega_0(1 + j/2Q_0) \quad .. \quad .. \quad .. \quad .. \quad (14)$$

In the perturbed condition, let Ω change from Ω_0 to $\Omega_0 + \delta\Omega$, with real part $\omega_0 + \delta\omega$, and let the Q now be Q_1. Then

$$\Omega_0 + \delta\Omega = (\omega_0 + \delta\omega)(1 + j/2Q_1) \quad .. \quad .. \quad (15)$$

Subtracting Equ. (14) from Equ. (15), and retaining small terms only of the first order, we obtain

$$\delta\Omega = \delta\omega + j\omega_0 \left\{ \frac{1}{2Q_1} - \frac{1}{2Q_0} \right\}$$

i.e., $$\frac{\delta\Omega}{\omega_0} = \frac{\delta\omega}{\omega_0} + j \left\{ \frac{1}{2Q_1} - \frac{1}{2Q_0} \right\} \quad .. \quad .. \quad (16)$$

Writing Equ. (13) in terms of the complex frequency, we obtain

$$\frac{\delta\Omega}{\omega_0} = -A \left\{ \frac{\epsilon' - j\epsilon'' - 1}{\epsilon' - j\epsilon'' - 2} \right\}$$

which becomes, for small ϵ'',

$$\frac{\delta\Omega}{\omega_0} = \frac{-A}{\epsilon' + 2} \left\{ \epsilon' - 1 - \frac{3j\epsilon''}{\epsilon' + 2} \right\}$$

Substituting for $\delta\Omega/\omega_0$ from Equ. (16), and equating real and imaginary parts, we finally obtain

$$\left.\begin{aligned} \frac{\delta\omega}{\omega_0} &= -A \left\{ \frac{\epsilon' - 1}{\epsilon' + 2} \right\} \\ \frac{1}{Q_1} - \frac{1}{Q_0} &= \frac{6A\epsilon''}{(\epsilon' + 2)^2} \end{aligned}\right\} \quad .. \quad .. \quad (17)$$

The first of Equs (17) shows that as long as ϵ'' is small the value of ϵ calculated from Equ. (13) is in fact the real part of ϵ. ϵ'' can be calculated from the second of Equs (17) if the Qs of the unperturbed and perturbed cavity are measured as well as the shift of resonance frequency.

While only a spherical sample has been considered here, analogous calculations can be made for samples of other shapes, both dielectrics and ferrites. For further information, references 3 and 7–11 may be consulted.

Dielectric Loss in a Simple Waveguide

The simple waveguide to be considered here consists of a metal tube of circular cross-section, containing air. Exact treatments of this guide were given in the second article of this series, using two models. For the first of these, the metal was taken to have the properties of a perfect conductor and the air those of vacuum; for the second, the metal was taken to have large but finite conductivity and the permeability of free space, and to extend to infinity, while the air was taken to have a dielectric constant $1 - j\epsilon''$, with $\epsilon'' \ll 1$. Using the second model, it was shown how the attenuation, due to dielectric losses in the air and conduction losses in the metal, could be calculated. We shall now calculate the attenuation due to dielectric loss in the air, regarding air as a perturbation of the vacuum of the first model,

and compare the perturbation result with the first-order result obtained by the use of the second model.

Equ. (4) gives the appropriate perturbation formula. The second bracket of the numerator of the right-hand side vanishes because there is no magnetic perturbation, so that $\mathbf{H}_1$ and $\mathbf{B}_1$ are zero. It was shown in an earlier section that $\mathbf{E}_1 = 0$ and $\mathbf{D}_1 = -j\epsilon''\epsilon_0\mathbf{E}_0$. Thus Equ. (4) becomes

$$\frac{\delta\omega}{\omega_0} = +j\epsilon''\epsilon_0 \frac{\iint_{S_s} E_0^2\, dS}{\iint_{S_s} (\epsilon_0 E_0^2 - \mu_0 H_0^2)\, dS} \quad .. \quad (18)$$

Now, the electric energy per unit length of the waveguide is

$$\frac{\epsilon_0}{2}\iint_{S_s} E_0^2\, dS$$

while the magnetic energy per unit length is

$$\frac{-\mu_0}{2}\iint_{S_s} H_0^2\, dS$$

Also, the electric and magnetic energies are equal. Thus the integral in the numerator of the right-hand side of Equ. (18) is half that in the denominator, and we have

$$\frac{\delta\omega}{\omega_0} = \frac{j\epsilon''}{2} \quad .. \quad .. \quad .. \quad .. \quad .. \quad (19)$$

It will be noticed that $\delta\omega/\omega_0$ depends only on ϵ'', not on the geometry of the waveguide nor on the mode of operation.

At constant frequency, it is $\delta\beta/\beta_0$ that is required, and this is obtained from Equ. (19) by the use of Equ. (5). It is well known that, for a simple waveguide of the type we are considering,

$$\frac{\omega}{\beta} = \frac{c}{\sqrt{1 - (\omega_c/\omega)^2}} \quad .. \quad .. \quad .. \quad .. \quad (20)$$

where ω_c is the cut-off value of ω. Differentiating,

$$\frac{d\beta}{d\omega} = \frac{1}{c\sqrt{1 - (\omega_c/\omega)^2}} \quad .. \quad .. \quad .. \quad (21)$$

Substituting from Equs (20) and (21) into Equ. (5), at frequency ω_0, we obtain

$$\frac{\delta\beta}{\beta_0} = \frac{-1}{1 - (\omega_c/\omega_0)^2}\frac{\delta\omega}{\omega_0}$$

and from Equ. (18) we obtain finally

$$\frac{\delta\beta}{\beta_0} = \frac{-j\epsilon''/2}{1 - (\omega_c/\omega_0)^2} \quad .. \quad .. \quad .. \quad .. \quad (22)$$

This gives $\delta\beta$ as negative and imaginary, i.e., $\delta\beta$ is an attenuation constant.

In the second article of this series, the result obtained by the result of a more sophisticated model was

$$\delta\nu = -j\epsilon''/2\nu_0 \quad .. \quad .. \quad .. \quad .. \quad .. \quad (23)$$

where $\nu = \lambda_0\beta/2\pi$, $\delta\nu = \lambda_0\delta\beta/2\pi$. From Equ. (20),

$$\nu_0 = \sqrt{1 - (\omega_c^2/\omega_0^2)}$$

Thus for Equ. (22) we may write

$$\frac{\delta\nu}{\nu_0} = \frac{-j\epsilon''/2}{\nu_0^2}$$

i.e., $\delta\nu = -j\epsilon''/2\nu_0 \quad .. \quad .. \quad .. \quad .. \quad (24)$

which is the same as Equ. (23). Thus perturbation theory gives the same result as the first-order expansion of the exact theory, and can be used if ν_0 is known. It may be, however, that ν_0 cannot be determined for certain modes from the model used, and then, of course, Equ. (24) is of no help. This is the case, for example, with the E_{p0} modes of the circular guide, discussed in the second article of this series.

Perturbation theory is only valid if $\delta\beta/\beta_0$ is small, so that β_0 must not be near zero. This applies generally, not only to the example treated above. Thus perturbation theory can only be applied to waveguides well away from cut-off, i.e. $\omega_0 - \omega_c$ must not be very small compared with ω_c.

Summary

Equs (1) and (2) give the frequency shift of a cavity due to perturbation. The use of these formulae enables the electric and magnetic properties of substances to be measured at microwave frequencies by observing the frequency shift when they are placed in a cavity.

Equs (3) and (4), coupled with Equ. (5), enable the shift of the phase constant of a waveguide to be calculated when a perturbation is made. Thus the attenuation constant, which may be regarded as a small imaginary part of the phase constant, can be calculated by regarding the metallic walls, with finite conductivity, as a perturbation of perfectly conducting walls, and the dielectric loss of the air in the guide as a perturbation of a lossless dielectric. A waveguide containing a rod of ferrite has been treated as a perturbation of a guide containing a rod of dielectric of the same permittivity and radius[12]; a thin rod of ferrite in a guide has been treated as a perturbation of an empty guide[12, 13], and a dielectric rod which nearly fills a waveguide has been treated as a perturbation of a guide filled with the dielectric[13]. In these last three examples, it is a real shift in phase constant that occurs. Other examples of the application of perturbation theory are the effect of a slight curvature[14], and the effect of pitch in a helical guide[15].

The possibility of using perturbation theory depends on the possibility of evaluating $\mathbf{E}_1$ and $\mathbf{H}_1$ in Equs (1), (2), (3), and (4). This may be difficult, but is often sufficiently simple to enable the theory to be used. It also depends on the solution's having been obtained of the characteristic equation for the model chosen. If the model does not give solutions for certain modes which are possible in the physical system, perturbation theory cannot be used.

In spite of these limitations, however, the examples cited show that perturbation theory is applicable to a wide range of important problems, particularly when only one mode is of interest, as is often the case in cavity measurements and waveguide applications. Where more than one mode is of interest, each mode must be treated separately, since $\mathbf{E}_1$ and $\mathbf{H}_1$ differ for the various modes. The approach through a more general model is often better in such cases, as an explicit expression for the first-order change of the eigenvalue can usually be obtained which is valid for all modes.

Acknowledgement

The author wishes to thank the Engineer-in-Chief of Marconi's Wireless Telegraph Company for permission to publish this article.

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Bandpass Transistor Amplifiers

ANALYSIS WITH DETERMINANTS

By C. J. McCluskey, B.Sc.*

SUMMARY. *In this article a determinant method is given for the analysis of the amplitude and phase responses of tuned amplifiers using transistors with complex internal feedback coupled by four- or two-terminal filter networks. Correction terms for a practical tuning procedure are presented. The form is well suited to electronic computation by which means multistage bandpass amplifiers may be dealt with.*

The design for a tuned transistor amplifier must usually take into account the feedback inherent to the device if instability or severe distortion of the amplitude/frequency response is to be avoided. The method often adopted is to determine the condition for oscillation and apply a stability factor obtained by analogy with some simple case such as a single stage. However, the condition for oscillation is not always known and extrapolation from simple cases is uncertain. In this analysis the transfer impedance of the amplifier as a whole is expressed in a determinant form containing the admittance parameters of the active and passive networks in simple terms. This form is well suited to electronic computation and is the intended method of application in all but simple cases. From it the effects of the internal feedback on the performance of multistage amplifiers can be determined. This determinant form was used by Jelonek and Sidorowicz[1] for the analysis of capacitive feedback in valve amplifiers with two-terminal coupling networks. That given here extends this to any system of cascaded four-poles, two-poles or mixed four- and two-poles. Correction terms for a practical tuning procedure in synchronous single and double-tuned amplifiers are presented. For applications where phase

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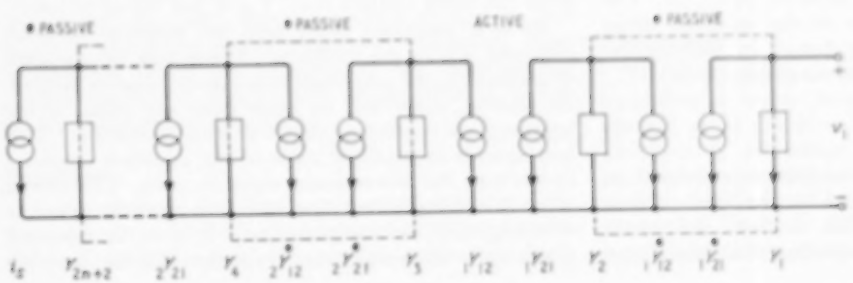

Fig. 1. *The equivalent circuit of an n-stage tuned transistor amplifier with four-terminal coupling networks for which 2n + 2 nodal equations may be written and solved by determinant methods*

is important, such as television i.f. amplifiers, the phase and envelope delay responses can be included in the computation.

Transfer Impedance—Determinant Form

Consider the following equivalent circuit of an n-stage tuned transistor amplifier with four-terminal coupling networks (Fig. 1). It is assumed that any transformations have already been made and incorporated into the lumped Y parameters. The asterisk is used to distinguish the transfer admittances of passive networks.

For this network $2n + 2$ nodal equations may be written and solved by determinant methods. The most important solution is the transfer impedance defined as the ratio of the output voltage to the driving-source current. It can be shown that this is given by

$$Z_{tn} = \frac{v_1}{i_g} = -\frac{\prod_{r=1}^{r=n} {}_rY_{21} \prod_{r=1}^{r=n+1} {}_rY_{21}^*}{\Delta_n} \quad \ldots (1)$$

where Δ_n is the determinant:

$$\begin{vmatrix} Y_{2n+2} & {}_{n+1}Y_{12}^* & {}_{n+1}Y_{21}^* & 0 & 0 \\ 1 & Y_{2n+1} & {}_nY_{12}\,{}_nY_{21} & 0 & \\ 0 & 1 & Y_{2n} & {}_nY_{12}^*\,{}_nY_{21}^* & \\ & & \ldots \text{etc.} \ldots & & \\ & & & Y_2 \;\; {}_1Y_{12}\,{}_1Y_{21} & 0 \\ & & & 1 \quad Y_2 \quad {}_1Y_{12}^*\,{}_1Y_{21}^* & \\ & & & 0 \quad 1 \quad Y_1 & \end{vmatrix}$$

The case of two-terminal coupling networks can readily be accommodated by compressing the determinant to order $n + 1$ by deleting the rows containing asterisk terms. The sign of the transfer impedance is then $(-1)^n$. Mixed two- and four-terminal networks could also be included. If numerical values are inserted in the determinant and it is multiplied out a real and imaginary part can be derived. Let these be written $\mathrm{Re}\Delta_n$ and $\mathrm{Im}\Delta_n$ respectively. From these the modulus and argument can be obtained to give the amplitude and phase responses respectively. The envelope delay can also be calculated in incremental form as part of the programme, viz.,

$$\underline{/Z_{tn}} = \phi = \pi + \sum_1^n \underline{/{}_rY_{21}} + \sum_1^{n+1} \underline{/{}_rY_{21}^*} + \arctan\left\{\frac{-\mathrm{Im}\Delta_n}{\mathrm{Re}\Delta_n}\right\} \quad \ldots (2)$$

In many cases ${}_rY_{21}$ and ${}_rY_{21}^*$ may be considered independent of frequency in the passband and it is sufficient to examine the last term in Equ. (2). The phase slope $d\phi/d\omega$, also known as the envelope delay or group delay, can then be found in incremental form by computing values of the function;

$$\frac{\Delta\phi}{\Delta\omega} \simeq \frac{\omega_0}{2Q}\frac{\Delta\phi}{\Delta x} \quad \ldots \ldots \ldots \ldots \ldots (3)$$

The transducer gain of an amplifier is defined as the ratio of the power delivered to the load to that available from the source. This form of the power gain has the advantage of combining the source and load conditions in a single homogenous expression. It is simply related to the transfer impedance by the equation,

$$\text{transducer gain } \Phi_{tn} = 4\,|Z_{tn}|^2\, G_S G_L \quad \ldots \ldots (4)$$

where G_S and G_L are the real parts of the source and load admittances respectively referred to the equivalent circuit given.

Regeneration Coefficients

Following Jelonek and Sidorowicz it is convenient to manipulate the terms of the determinant to illustrate the factors that control the stability. As an example consider a single-stage double-tuned amplifier. The admittance parameters for the equivalent circuit may be written:

$$Y_r = G_r\{1 + j(x + x_r)\}, \qquad {}_rY_{12}^* = {}_rY_{21}^* = -j(KQ)_r\, G_{r+1}\, G_r$$

$$\text{and} \quad Y_{12}\, Y_{21} = Me^{j\theta}$$

Both M and θ are to be taken positive to avoid ambiguity. (For example, for purely capacitive feedback θ is 270°). Then the determinant may be written

$$\Delta_1 = G_1G_2G_3G_4 \begin{vmatrix} 1+j(x+x_4) & -(KQ)_2{}^2 & 0 & 0 \\ 1 & 1+j(x+x_3) & \dfrac{Me^{j\theta}}{G_3G_2} & 0 \\ 0 & 0 & 1+j(x+x_2) & -(KQ)_1{}^2 \\ 0 & 0 & 1 & 1+j(x+x_1) \end{vmatrix} \quad (5)$$

The x terms represent the fractional detunings of the circuits. If the circuits are individually tuned to the same frequency then all of the x_r terms are zero. The coefficient $M/(G_3G_2)$ will be called, after Jelonek and Sidorowicz, the 'regeneration coefficient'. The condition for oscillation in the amplifier is that the determinant becomes zero at some frequency. (Both real and imaginary parts simultaneously.) For given tuned circuits and a given value of θ the value of the regeneration coefficient determines whether this will happen or not. If the transistor is perfectly neutralized θ is zero and oscillation is not possible.

However, component tolerances, transistor-parameter spreads and variations in parameters, due to change of operating point with a.g.c., may make it necessary to consider a value for θ under practical conditions. In this and unneutralized cases the stability is then controlled by the value of the regeneration coefficient. If it is arbitrarily put equal to zero the expressions represent the performance of a unilateral amplifier with the same circuits (i.e., Y_{12} is put equal to zero).

As it is increased from zero the response becomes distorted due to the feedback. Eventually, at the point of oscillation, the amplitude response becomes unlimited. In general, there will be more than one value of the regeneration coefficient that will give oscillation. This corresponds to multiple roots of the equation obtained by putting the determinant equal to zero. The lowest value is the significant one and care must be taken in selecting data for computation if it is to be detected correctly. However, the response will be severely

affected even with values only half of this so the point of oscillation is largely of academic interest except as a guide to setting up data for the programmes.

Tuning Corrections

If the admittance parameters of the individual circuits making up the amplifier are written into the determinant directly, they imply a tuning procedure whereby each circuit is tuned with all the others heavily damped. This results in an appreciable shift of the overall response from the nominal centre frequency if the feedback is reactive and in any case does not represent the normal tuning procedure. A more practical approach is to connect a signal generator, terminated with an impedance low enough to damp the circuits, across each circuit in turn beginning at the last stage. The following circuit in each case is tuned to give maximum output from the whole amplifier at the required centre frequency.

Mathematically, this procedure may be analysed by considering the transfer impedances relative to the generator at each step. The expressions for these contain determinants which are parts of the main determinant, the order increasing by one at each step until ultimately the complete determinant is obtained. It is found that for maximum output at the centre frequency the tuning terms are adjusted so that the partial determinants are wholly real. These correction terms may be identified with the terms x_r given earlier. Different considerations apply to amplifiers based on stagger-tuned filters. For equal synchronous single-tuned circuits with identical transistors so that the regeneration coefficients have a common value T and the feedback a common phase angle θ, the correction terms are as follows

$$x_1 = 0,\quad x_2 = T\sin\theta,\quad x_3 = T\sin\theta\,\frac{1}{(1 - T\cos\theta)},$$

$$x_4 = T\sin\theta\,\frac{(1 - T\cos\theta)}{(1 - 2T\cos\theta)},$$

$$x_5 = \frac{T\sin\theta\,(1 - 2\,T\cos\theta)}{(1 - 3T\cos\theta + T^2\cos^2\theta)},\ \text{etc.}$$

For equal synchronous double-tuned circuits the correction terms may be expressed by iterative formulae distinguishing between even and odd terms, viz.,

$$x_{2r} = 0,\ x_{2r+1} = T\sin\theta\,\frac{P_{2r-2}}{P_{2r-1}},\quad x_1 = 0,\quad P_0 = 1,$$

$$P_1 = 1 + (KQ)_1{}^2,\quad P_{2r} = P_{2r-1} - T\cos\theta\,.\,P_{2r-2},$$

$$P_{2r-1} = P_{2r-2} + (KQ)_r{}^2\,.\,P_{2r-3}.$$

The relative complexity of these terms is of little consequence when electronic computation is employed. It is to be noted that these corrections do not imply nor do they generally result in a response that is symmetrical about the centre frequency. However, they appreciably affect the response in the direction of greater stability so it is usually worth while to incorporate them in the calculation.

Example

As an example, consider a two-stage amplifier with three equal synchronous double-tuned coupling circuits $KQ = 1$, and $\theta = 270°$ (i.e., purely capacitive feedback). It is known that an isolated stage of the amplifier would oscillate if the regeneration coefficient were about 3·2. Therefore the values 0, 0·5, 1·0, 1·5, and 2·0 are chosen. From the known unilateral response x is chosen to have the values −2·0 to +2·0 in increments of 0·2. The appropriate tuning corrections were included. The results of the computation were plotted as amplitude and phase slope responses as shown in Fig. 2.

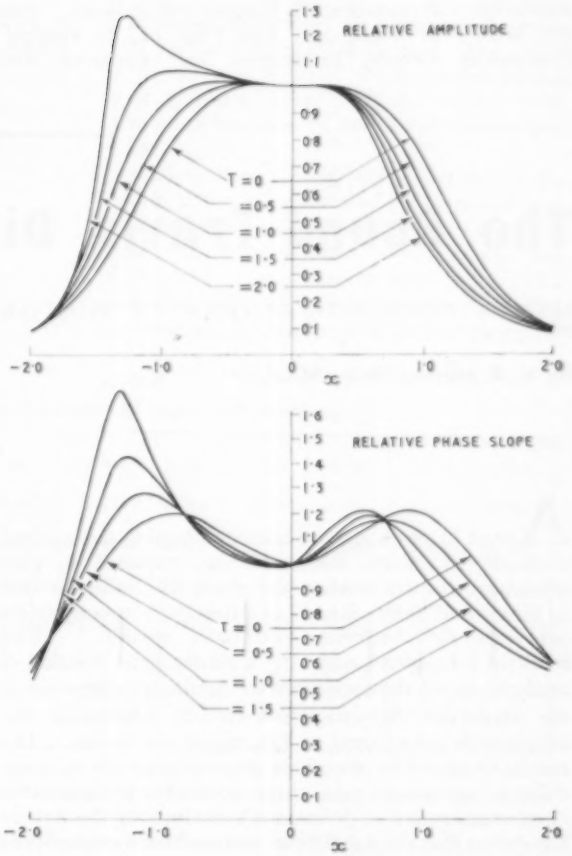

Fig. 2. *Taking as an example a two-stage amplifier with three equal synchronous double-tuned coupling circuits the amplitude and phase-slope responses against x were computed and plotted*

Conclusion

The determinant method of analysing tuned transistor amplifiers presented here enables the designer to assess the effects of complex internal feedback on the performance of multistage bandpass amplifiers with the aid of electronic computation. From this a satisfactory design centre may be chosen and the corresponding theoretical limit to the gain established. This also assists the designer in minimizing external feedback

due to construction techniques by setting a limit to the improvements theoretically possible.

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The Round-Travis Discriminator

MINIMIZING NON-LINEARITY DISTORTION

By K. R. Sturley, Ph.D., M.I.E.E.*

About 15 years ago the author, adopting a graphical method[1], examined the conditions required to give minimum distortion when the phase discriminator due to Foster and Seeley is used as a frequency-to-amplitude converter for frequency-modulated signals. More recently Johnstone[2], adopting a power series method of analysis, found the conditions for minimum distortion in the amplitude (Round-Travis circuit) and phase discriminators when used as f.m.-signal converters. The results obtained for the phase discriminator are substantially in agreement with those determined graphically. A re-examination of Johnstone's analysis by the author has shown that the conditions indicated for the amplitude discriminator were not those giving minimum distortion, and the purpose of this note is to add a correction.

Johnstone quotes as the expression for the output voltage for the amplitude discriminator (Fig. 1):

$$V_{out} = K_1[\{1 + (x + x_1)^2\}^{-\frac{1}{2}} - \{1 + (x - x_1)^2\}^{-\frac{1}{2}}] \quad (1)$$

where K_1 = constant

$$x_1 = \frac{Q2\Delta f_1}{f_0}$$

$$x = \frac{Q2\Delta f}{f_0}$$

Q = Q factor of circuits 1 and 2 (assumed equal),

$\Delta f_1 = f_1 - f_0 = f_0 - f_2$

f_1 = resonance frequency of circuit 1,

f_2 = resonance frequency of circuit 2,

f_0 = mid-frequency = $\frac{1}{2}(f_1 + f_2)$

Δf = the variable, off-tune frequency from f_0

Equ. (1) may be rewritten as a power series

$$V_{out} = K_2(a_1x + a_3x^3 + a_5x^5 + \ldots\ldots) \quad \ldots \quad (2)$$

whose coefficients, specified by Maclaurin's theorem, are

$$a_1 = 2x_1(1 + x_1^2)^{-3/2}$$

$$a_3 = x_1(2x_1^2 - 3)\ (1 + x_1^2)^{-7/2}$$

$$a_5 = \tfrac{1}{4}\, x_1(8x_1^4 - 40x_1^2 + 15)\ (1 + x^2)^{-11/2}$$

The modulation content of the f.m. signal is contained in the factor x which is proportional to the deviation frequency $f_d \cos 2\pi f_m t$, so that a_3x^3 produces 3rd harmonic distortion of the modulation and a_5x^5 produces 3rd and 5th harmonics. Johnstone suggests that mini-

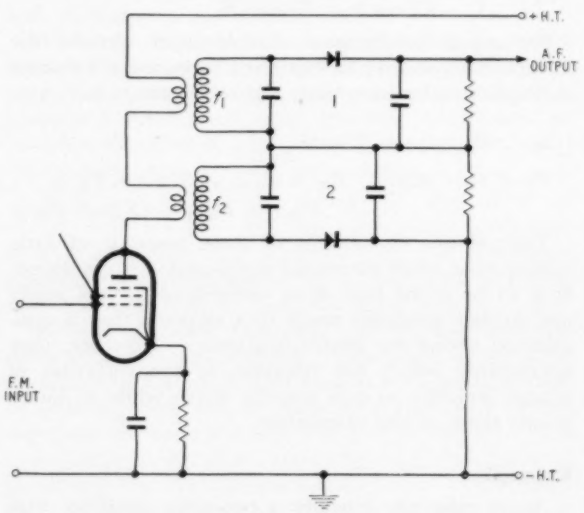

Fig. 1. *The amplitude discriminator as f.m. converter*

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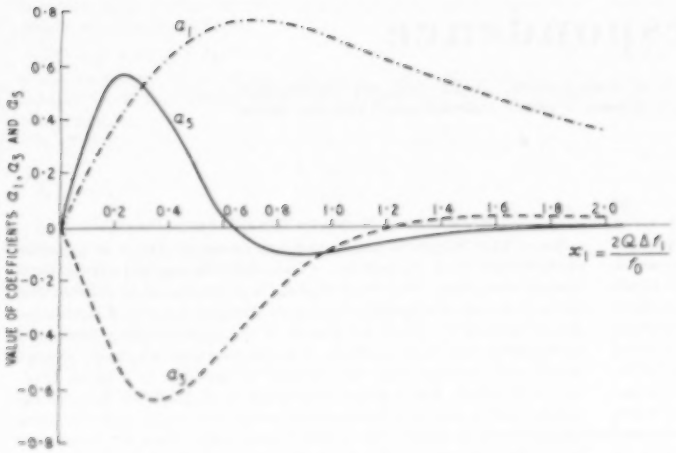

Fig. 2. *The power-series coefficients for the amplitude discriminator plotted against* $x = \left(\frac{2Q\Delta f_1}{f_0}\right)$

mum modulation distortion is achieved when $a_3 = 0$, which gives $x_1 = \sqrt{1{\cdot}5} = 1{\cdot}225$.

If the coefficients a_1, a_3 and a_5 are plotted against x_1, as in Fig. 2, it is seen that over a certain range of x_1, a_3 is positive when a_5 is negative and this suggests that 3rd harmonic distortion due to the a_3 term may be made to cancel the 3rd-harmonic distortion produced by the a_5 term. The value of the distortion content of the a_3 term arises from

$$a_3x^3\cos^3\theta = a_3x^3(3\cos\theta + \cos 3\theta)/4$$

and that from the a_5 term is

$$a_5x^5\cos^5\theta = a_5x^5(10\cos\theta + 5\cos 3\theta + \cos 5\theta)/16$$

Thus for zero 3rd harmonic

$$a_3 = 5a_5x^2/4 = 1{\cdot}25a_5x^2 \quad .. \quad .. \quad .. \quad (3)$$

The above condition can only be met at a single value of x, so let it be assumed that this occurs when $x = 1$ which corresponds to a deviation frequency $f_d = 75$ kc/s. Solving Equ. (3) gives $x_1 = 1{\cdot}44$, and the table below gives the values of a_1, a_3 and a_5 as well as the percentage 3rd and 5th-harmonic distortion for $x = 0{\cdot}5$, 0·7 and 1, when $x_1 = 1{\cdot}225$ (Johnstone's minimum), 1·4, 1·44 and 1·5. The percentage harmonic distortion is calculated from $100(a_3x^2 + 1{\cdot}25\,a_5\,x^4)/4a_1\%$ and $100\,a_5\,x^4/16\,a_1\%$ the contribution to the fundamental from a_3 and a_5 being neglected because both are much less than a_1 and x does not exceed unity.

Clearly $x_1 = 1{\cdot}4$ represents the best compromise; there is a reduced fundamental output compared with that at $x_1 = 1{\cdot}225$, but distortion is very much reduced and less dependent on modulation amplitude. The required value of Q is determined by x and is therefore the same as that calculated by Johnstone; viz., 71, for a mid-frequency of 10·7 Mc/s but each circuit must be off-tuned by $1{\cdot}4f_0/2Q = 105{\cdot}5$ kc/s from the mid frequency 10·7 Mc/s.

TABLE

Percentage Distortion in terms of Input Signal

x_1	a_1	a_3	a_5	%3rd	%5th	%3rd	%5th	%3rd	%5th
1·225	+0·62	0	−0·055	0·17	0·034	0·65	0·13	2·7	0·54
1·4	+0·55	+0·0289	−0·03	0·22	0·021	0·23	0·082	0·39	0·342
1·44	+0·537	+0·032	−0·025	0·28	0·018	0·37	0·068	0	0·285
1·5	+0·513	+0·036	−0·018	0·365	0·0135	0·6	0·052	0·66	0·205
x				0·5	0·5	0·7	0·7	1·0	1·0
f_d				37·5kc/s		52·5kc/s		75 kc/s	

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Letters to the Editor on technical subjects are always welcome. In publishing such communications the Editors do not necessarily endorse any technical or general statements which they may contain.

Transistor Circuit Diagrams

SIR,—In the early days of the transistor art, when the physicist held sway, it was decreed that the transistor was the dual of the valve and that valve engineers should, henceforth, stand on their heads or go out of business. A survey of electronic-engineering periodicals shows that this assessment was incorrect and that engineers prefer that their transistor circuit should look like valve circuits, a trend towards the unified theory previously discussed in your columns. However, the Standards Organizations and the drawing offices are now entering the field and are suggesting that we must, at all costs, abide by a standard that requires the top h.t. rail to be positive and the bottom h.t. rail to be negative with all intervening circuits to be upside down, if p-n-p transistors are used, and the right way up if valves or n-p-n transistors are used.

Personally, I am unwilling that the vast accumulation of background knowledge and ready appreciation of a circuit function gathered during several hundred-thousand man-years, I write in a collective sense, of experience in radio and electronics generally should be jettisoned in order to satisfy the early 1950 conceptions of the physicists. The simple point to grasp is that an emitter is a cathode, a base is a grid and a collector is an anode, a point made previously in your columns.

Furthermore, provided that the topography of a circuit, valve or transistor, is unchanged, then there will be less confusion when n-p-n transistors are generally used. We must consider the effect of this varying topography upon the service engineer and technician. If we can hammer the point that these circuits are the circuits that he already knows, with minor exceptions, then we shall achieve a considerable advance in general transistor knowledge, something lacking at the moment due to the esoteric nature of early writing on the subject.

It may be argued that in mixed (complementary) circuits the foregoing arguments fail, in fact they do not. In such circuits a certain mental agility is necessary and, if understanding is difficult, more practice in the art is necessary. As a general principle, the layout of the transistor circuit should be dictated by the predominating transistor type.

It may be argued that, if the negative h.t. line is at the top of the circuit, interpretation of the c.r.t. waveform will be inhibited. If understanding of a circuit function is trammelled by the h.t. line then, perhaps, the h.t. line should be removed. To suggest that the circuit should be inverted is, to my mind, akin to a suggestion that we should adopt back-slang as a technical language.

In this laboratory we are dealing simultaneously with the design and development of both valve and transistor circuits and the engineers are accustomed to changing rapidly from one to the other; there is no difficulty in waveform interpretation, the alleged mental blocks having been readily overcome. However, when discussing a circuit it is found advantageous to indicate the nature of the waveform by means of a plus or minus symbol appended to the peak of the waveform. For example, within a p-n-p circuit, with negative h.t. at the top, the waveforms that move towards the h.t. line are drawn moving upwards and have a negative sign appended adjacent to the peak. It is found that this convention enables discussion to carry on without difficulty in hybrid, complementary or valve circuits.

There does seem to be a real problem arising in connection with Standards of Drawing that may be resolved by discussion among engineers. At the moment I find that engineers "on the bench" are drawing their circuits the "right way up" and tamely accepting the drawing office reversal of the drawing. If this is general we may find that a British Standard may be issued contrary to engineering practice but in conformity with drawing office practice.

T. G. CLARK.

Radar Development Laboratory,
Decca Radar Ltd.,
Chessington, Surrey.
29th March 1961.

Barkhausen Noise in Transformer Cores

SIR,—Mr. Warren's paper (March issue, p. 87) is a valuable contribution to a subject on which there is comparatively little quantitative data. The result that e_n^2 is proportional to dH/dt is due to the fact that the Barkhausen jumps produce a series of randomly-spaced impulses. Since the phases of the corresponding frequency components are then random, it is the mean-square noise voltage which will increase with the number of impulses in a given time; i.e., with dH/dt. For a given total range in B, $\Sigma\Delta B$ will be constant (where ΔB is due to a Barkhausen jump) and under this condition $\Sigma(\Delta B)^2$ will be larger for a few coarse steps than for many fine steps. But, in general, the permeability is greater if the Barkhausen jumps are large, so that $\Sigma(\Delta B)^2$ is proportional to the permeability; i.e., to the mean of dB/dH.

So far as the spectrum is concerned, it would be interesting to have the data of Fig. 12 re-plotted on logarithmic scales, so as to make it clear whether there is a constant power law. A point to watch is the effect of eddy currents on the external field resulting from a change in a small domain within the sample. Tebble et al have calculated this effect for round wire[1], but I do not know of any calculation for plane laminations. However, one would not expect eddy-current effects to be appreciable at the frequencies below 5 c/s to which Fig. 10 refers.

In Fig. 11 the author suggests that the pulses have an exponential shape, and eddy currents will probably ensure that this is so. The corresponding power spectrum is then of the 'relaxation' form,

$$W(f) \propto \frac{\tau}{1 + 4\pi^2 f^2 \tau^2}$$

so that the mean-square noise voltage eventually falls as the square of the frequency, and it is the r.m.s. noise voltage $\sqrt{e_n^2}$ which decays inversely as the frequency for values in excess of $1/\tau$.

Electrical Engineering Department, D. A. BELL.
University of Birmingham,
Edgbaston, Birmingham, 15.
16th March 1961.

REFERENCE

[1] R. S. Tebble, I. C. Skidmore, and D. W. Comer, "The Barkhausen Effect". *Proc. Phys. Soc. A.*, 1950, Vol. 63, p. 739.

Non-Resonant End-Fed Aerials

SIR,—In connection with Dr. Talekar's paper in your January issue, may I suggest an exact formulation for θ_m, the angle of maximum radiation? This is

$$\theta_m = \cos^{-1}\left[1 - \frac{\lambda}{2l} + \frac{\lambda}{\pi^2 l}\cdot\frac{F(\lambda/l)}{1-\lambda/4l}\right];\ 0{\cdot}46 < \frac{l}{\lambda} < \infty$$

$$\theta_m = \cos^{-1}\left[\frac{\pi^2 l^2}{3\lambda^2}\,G(\lambda/l)\right];\ 0 \leqslant \frac{l}{\lambda} \leqslant \frac{1}{\pi}$$

where $F(\lambda/l)$ and $G(\lambda/l)$ are convergent series,

$$F(\lambda/l) \equiv 1 + \frac{5/3-\lambda/l}{\pi^2(1-\lambda/4l)^2} + \frac{2(5/3-\lambda/l)^2 + (1-\lambda/4l)\lambda/l}{\pi^4(1-\lambda/4l)^4} + \ldots.$$

$$G(\lambda/l) \equiv 1 - \frac{2}{3}\frac{\pi^2 l^2}{\lambda^2} + \left[\frac{5}{9} + \frac{2}{5}\frac{\lambda^2}{\pi^2 l^2}\right]\frac{\pi^4 l^4}{\lambda^4} + \ldots\ldots$$

Over a small region where l/λ lies between $1/\pi$ and $0{\cdot}46$ the solution is indeterminate because the polar diagram there exhibits no simple (unique) maximum.

These relations are derived from Talekar's equation (8)

$$\tan x = \left(2 - \frac{x\lambda}{\pi l}\right)x \qquad \ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

where $x = \dfrac{\pi l}{\lambda}(1 - \cos\theta)$

The angle of maximum radiation θ_m is the solution of this equation.

If we have an equation

$$y = a_1 X + a_2 X^2 + a_3 X^3 \ldots\ldots \qquad \ldots\ldots\ldots\ldots \quad (2)$$

in which $a_1 \neq 0$ and the r.h.s. is a convergent series, there is a unique solution in the form

$$X = b_1 y + b_2 y^2 + b_3 y^3 \ldots\ldots \quad (3)$$

which converges in a certain interval[1]. The coefficients b_n are formally found by substituting (3) into (2) and equating like powers of y, giving

$$b_1 = 1/a_1; \qquad b_4 = -(5a_2^3 - 6a_1a_2a_3 + a_1^2a_4)/a_1^7;$$

$$b_2 = -a_2/a_1^3; \qquad b_n = \frac{1}{n!}\frac{d^{n-1}}{dX^{n-1}}\left(\frac{1}{\psi(X)}\right)^n X = 0$$

$$b_3 = (2a_2^2 - a_1a_3)/a_1^5; \quad \psi(X) = a_1 + a_2X + a_3X^2 \ldots\ldots$$

Now the range of interest of x in equation (1) is 0 to π. The variable x is changed to X where $x = \pi/2 - X$ and equation (1) becomes

$$0 = \cot X - (\pi/2 - X)\,[2 - (\pi/2 - X)\,\lambda/\pi l] \quad (4)$$

Using the series expansion

$$\cot X = 1/X - X/3 - X^3/45 \ldots\ldots$$

equation (4) becomes

$$1 = \pi\,(1 - \lambda/4l)X - (5/3 - \lambda/l)X^2 - X^3\,\lambda/\pi l + X^4/45 \quad (5)$$

This is of the form of equation (2) and so the coefficients of (3) can be found from the relations given. Returning to the original variable x and simplifying, we get the first expression for θ_m. The second is obtained in the same way but without a change of variable, the series expansion for tan x being substituted directly in (1).

As always with series, it is necessary to check that they are convergent. Space does not permit the inclusion of the details of the tests applied but they show that the relations given at the start of this letter are valid except for the range of values of l/λ between $1/\pi$ and 0·46.

J. Deignan

Engineering Division,
Radio Eireann,
Athlone, Eire.

1st March 1961.

REFERENCE

[1] Goursat and Hedrich, "A Course in Mathematical Analysis", Vol. 1, Section 190.

New Books

Introduction to Modern Network Synthesis
By M. E. Van Valkenburg. Pp. 498 + xii. Chapman & Hall (for John Wiley & Sons), 37 Essex Street, London, W.C.2. Price 94s.

This book represents a further attempt to elucidate a vital but rather difficult subject. During recent years, attention has been paid on an increasing scale to the exposition of network synthesis. This is good, for so long as a considerable quantity of literature has remained uncorrelated and unsimplified, techniques for the design of networks on a universal and explicit basis, rather than on a quasi-empirical basis, have been accessible to but few.

Notable contributions to the consolidation of synthesis principles and techniques have already been made through the books of E. A. Guillemin ("Synthesis of Passive Networks"), D. F. Tuttle ("Network Synthesis"), and R. C. Storer ("Passive Network Synthesis"). Professor Van Valkenburg's book, while following more or less an established pattern, represents to some extent a compromise between the brevity of Dr. Storer's excellent survey of principles and methods, and the depth and detail embodied in the other works mentioned. These works are, in fact, frequently cited as references, and tribute is paid to Dr. Tuttle in the preface.

An introductory chapter, in which an attempt is made to define the synthesis problem and to illustrate the significance of approximation, includes a brief but nevertheless lucid development of the complex-frequency and pole-zero concepts. Tabulated diagrams are used with advantage to identify real, imaginary and complex frequencies, and pole locations, with familiar forms of steady-state and transient (or "time-domain") responses.

The use of "L.L.F.P.B." as an abbreviation for "lumped, linear, finite, passive, and bilateral" is, however, somewhat pedantic. For a reader competent to use the book, repetition of this description is but rarely occasioned; and it is most undesirable that the fashion for initials, which in other spheres has grown to such proportions that a special dictionary is almost warranted, should be allowed to complicate further a subject that is inherently complicated enough.

After defining analysis and synthesis in terms of the network, excitation, and response, the author says: "The third possibility, given the network and the response and required to find the excitation, has no generally accepted name and is not common". It is agreed that in telecommunications this permutation is encountered less frequently than those of analysis and synthesis; but in the vast field of electric power transmission, it is not merely common, it is fundamental: the specified quantity is normally the load kVA, etc. Moreover, it is noteworthy that the order of terms in equations of the type $V_1 = AV_2 + BI_2$, $I_1 = CV_2 + DI_2$, is allied to the fundamental nature of power-transmission problems, to which at one time such equations were almost exclusive. Although this order is accepted as standard, the order $V_2 = AV_1 + BI_1$, $I_2 = CV_1 + DI_1$ might seem more appropriate to network analysis, in which the response (V_2, I_2) is to be found as a function of the excitation (V_1, I_1).

It is interesting to learn from a footnote on page 20 that "*Another* frequency unit is the cycle per second, denoted by the symbol *f* and related to ω by a 2π scale factor, $\omega = 2\pi f$." Apparently (on the same page) ω has come to be known as real *frequency*, although the authors adds: "Another *convention* is to name ω radian frequency—."

The fifteen chapters that follow are divided into three groups by the topic of approximation, which is treated in two phases in the ninth and thirteenth chapters. The first group covers: Network analysis; Positive-real Functions; Testing; Driving-point Synthesis, with reference to both pure reactance and dissipative one terminal-pair networks; and Parts of Network Functions. The second group, which follows a discussion of general features and methods of approximation in Chapter 9, deals with the synthesis of two terminal-pair networks by Ladder Development and by Series and Parallel Realisation with ideal transformers, and with the lattice network. The second phase of the approximation topic, in Chapter 13, is devoted to optimization and the Butterworth and Chebyshev responses. The final chapters deal with terminated networks, Darlington's approach, and the image-parameter method, including frequency transformations. Each chapter is accompanied by the customary list of problems without answers although, on the other hand, there are many worked-out illustrations in the text.

The compact chapter on analysis is useful, for the procedures of analysis and synthesis are, with reservations, complementary, and the immittance parameters are common to both. The author's derivation of transfer voltage-ratio for a two-terminal-pair network, based on the theorems of both Norton and Thévenin, is however, unnecessarily involved. A simple derivation is possible with Thévenin's theorem alone, and the result is also implicit in the nodal equations.

The presentation as a whole has many good features. Lucidity has been enhanced by the omission of non-essential mathematical details, which are available in the references quoted and in the lists given for further reading. In general, the author has succeeded in bringing out the essence of synthesis procedures. This is exemplified, in respect of driving-point synthesis, on p. 114; by the explanation of odd and even functions (p. 97); in the comparison of *LC* with *RC* networks (p. 142); by the practical attitude to approximation; and in the brief but very clear explanation of the potential analogy on p. 234.

This book goes somewhat further, perhaps, in giving synthesis principles a practical interpretation. After examining the books on this topic by a number of authors, one cannot ignore the impression that, so far, most have written either as mathematicians having little contact with the practical problems and objectives of professional engineers, or in the false belief that extensive illustration with concrete practical problems detracts from the status traditionally connected with a highly academic approach. The division between the academic and practical interests is in fact summarized by Prof. Van Valkenburg when he says in his preface: "To some, network

synthesis is strictly applied mathematics; to others, it is a means for solving day-by-day problems". With reference to current books on synthesis, it has been said by the head of a filter-design department: "Life is too short".

Those who write books on synthesis in the future will have little to offer if they confine themselves to repetition of the mathematical background and procedures already thoroughly explained in the works of E. A. Guillemin, D. F. Tuttle and others. They might well consider objectives such as: simplification in the presentation of basic principles; the sifting and appraisal of the various approaches, in relation to the classes of problem encountered in practice; and, particularly, the crystallization of procedures through extensive numerical and descriptive illustrations based on practical problems. The engineer, whose practical objective is no less creditable than the theoretical one of the mathematician, is in need of clear-cut and systematized data on the synthesis of networks.

Professor Van Valkenburg's book is very well produced, and is recommended particularly for the lucidity with which it presents the subject. F.E.R.

"Wireless and Electrical Trader" Year Book 1961

Compiled in collaboration with the staff of *Wireless and Electrical Trader*. Pp. 447. Published by Iliffe Books Ltd., Dorset House, Stamford Street, London, S.E.1. Price 15s. (12s. for subscribers to *Wireless and Electrical Trader*).

Listed as "a legal, technical and buying guide for the radio, television and domestic electrical industries" this 32nd edition of the well-known year book contains twenty-two main sections. Some of these are: Directory of Principal Trade Organizations, Legal Guide, Field-Strength Contour Maps, Valve-Base Connections, Cathode-Ray Tube Data, Transistor Types and Connections, Wholesalers, Proprietary Names Directory, Buyers' Guide and Trade Addresses.

Noble Metal Thermocouples (3rd Edition)

By H. E. Bennett, F.I.M. Pp. 43. Published by Johnson Matthey & Co. Ltd., 73–78 Hatton Garden, London, E.C.1. Available free on application.

"This book describes the development of the noble metal thermocouple, some of its applications and methods of calibration. Curves of temperature against e.m.f. are given for many types of couple incorporating alloys of the platinum group metals and a chapter describes the chief causes of deterioration in service.

An addition to the present edition is a short section summarizing the behaviour of many thermocouple combinations and recommending those most suitable for particular applications."

Handbook of Noise Measurement (4th Edition)

By Arnold P. G. Peterson and Ervin E. Gross, Jr. Pp. 132. Published by General Radio Company, West Concord, Massachusetts, U.S.A. Price $1.00.

"In thirteen chapters, this edition offers a comprehensive survey of such topics as the decibel, man as a noise-measuring instrument, applications of sound-measurement systems and the measurement of sound levels and sound-pressure levels. Also detailed are the characteristics of noise control, loudness levels, speech interference, hearing damage and neighbourhood reaction to noise."

Translation from Russian for Scientists

By C. R. Buxton and H. S. Jackson. Pp. 299 + xix. Blackie & Son Ltd., 17 Stanhope Street, Glasgow, C.4. Price 30s.

"This book is designed primarily to enable the scientist to acquire an adequate reading knowledge of the Russian language. He is first introduced to the alphabet by means of scientific words that are similar to their English counterparts, and from there he proceeds to essential grammar and translation passages taken from various scientific spheres."

The Relay Guide

By Raymond N. Auger. Pp. 360 + xv. Published by Reinhold Publishing Corporation, New York and distributed in the U.K. by Chapman & Hall Ltd., 37 Essex Street, London, W.C.2. Price 80s.

"The goal of this book is a thorough description of every electromechanical relay being marketed in the United States. The editors have transcribed information from hundreds of catalogs and bulletins into a relatively standardized format to help the relay user find the item which seems best suited to his requirements. Because our information has been obtained from literature sources, there is considerable variation in the thoroughness of the descriptions."

Pergamon Press Catalogue of Scientific and Technical Publications: 1961

Pp. 226. Published by Pergamon Press, Headington Hill Hall, Oxford. Available on request.

"A complete listing of all Pergamon publications currently in print or in press as of 1 December 1960."

On Microwave Conductivity, Noise, and Oscillations of Gas Discharge Plasma

By Paaro Jaaskelainen. Pp. 22. Published by the State Institute for Technical Research, Helsinki, and available from Acta Polytechnica Scandinavica Publishing Office, Box 5073, Stockholm 5. Price Sw. Kr. 7.00.

The Arc Discharge: Its Application to Power Control

By H. de B. Knight, M.Sc., M.I.E.E., F.Inst.P. Pp. 444 + xix. Chapman & Hall, 37 Essex Street, London, W.C.2. Price 63s.

"This book is concerned with the use of arc-discharge valves such as thyratrons, ignitrons and excitrons for the control of electrical power. It describes the construction and operating characteristics of the different types available; their capabilities and limitations; and their behaviour in miscellaneous rectification and control circuits.

The physical processes involved inside the valves are also discussed; the aim here is to show the basis of rating limitations specified by the manufacturers and also give some guidance as to the behaviour to be expected in new applications not covered by the ratings."

Tube and Semiconductor Selection Guide: 1960–61

By Th. J. Kroes. Pp. 157. Published by Philips' Technical Library and distributed in the U.K. by Cleaver-Hume Press Ltd., 31 Wright's Lane, London, W.8. Price 12s. 6d.

Selection and Application of Semiconductor Devices: ARINC Specification No. 409

Pp. 50. Published by Aeronautical Radio Inc., 1700 K Street, N.W., Washington 6, D.C., U.S.A. Price $1·00.

". . . a truly international preferred list, co-ordinated by the Airlines Electronic Engineering Committee (U.S.A.) through the European Airlines Electronic Committee, based on transistors described by various U.S., British and Dutch military and commercial specifications."

Note: This issue is limited to transistors.

NATIONAL BUREAU OF STANDARDS

Standardization Activities in the United States

By Sherman F. Booth. Pp. 210. National Bureau of Standards Miscellaneous Publication 230. Price $1·75.

"This directory provides a descriptive inventory of the work and objectivities of organizations, both public and private, involved in standardization activities. Each organization is listed alphabetically with a brief description of its standardization area and accomplishments."

Mechanical Properties of Structural Materials at Low Temperatures: A Compilation from the Literature

By R. M. McClintock and H. P. Gibbons. Pp. 180. National Bureau of Standards Monograph 13. Price $1·50.

"Intelligent design of reliable cryogenic equipment requires data on the mechanical properties of structural solids at low temperatures. This Monograph is issued to help fill the need for a compilation of useful design figures. It includes tensile strength, yield strength, tensile elongation, and impact energy of about 200 materials, metallic and non-metallic, given graphically as functions of temperature between 4 to 300°K."

Handbook "Preferred Circuits", Navy Electronic Equipment NAVWEPS 16.1.519, Supplement No. 3

Pp. 100. Published by the National Bureau of Standards. Price $0·55.

"This latest supplement to "Handbook Preferred Circuits"

contains seven germanium transistor circuits (PC210 through 216), two silicon transistor circuits (PC252 and 253), and one vacuum-tube instrument servo circuit (PC79). Preferred circuits 210–216 are a compatible set of transistor digital logic circuits for use in computer, control, and communication equipment. PC252 is a variable-gate generator, and PC253 is a non-saturating bistable multivibrator capable of operating at rates up to 1 Mc/s. The vacuum-tube circuit is a phase-sensitive null-detector for use in servo control circuits."

Precision Measurement and Calibration (Vols. 1, 2 and 3)

Vol. 1, "Electricity and Electronics". Pp. 845. Price $6.00; Vol. 2, "Heat and Mechanics". Pp. 965. Price $6.75; Vol. 3, "Optics, Metrology and Radiation". Pp. 1,025. Price $7.00.

"This Handbook is a compilation of the more important National Bureau of Standards' publications over a period of years, dealing with precision measurement and the calibration of standards."

Report of the International Commission on Radiological Units and Measurements

Pp. 90. Price 65 cents (postage 17 cents).

Research Highlights of National Bureau of Standards: Annual Report 1960

Miscellaneous publication No. 237. Pp. 189. Price 65 cents.

These publications are available from the Superintendent of Documents, U.S. Government Printing Office, Washington 25, D.C., U.S.A.

Remittances should include an additional one-fourth of the publication price to cover mailing cost.

BRITISH STANDARDS

Glossary of Terms used in Telecommunication (including radio) and Electronics

Pp. 350. B.S. 204: 1960. British Standards Institution, 2 Park Street, London, W.1. Price 35s.

"The 2,700 definitions in this new volume represent a decade's work on the part of one of B.S.I.'s many committees of experts. The standard brings together all the terms previously dealt with in the 1943 edition and the various supplements published between 1948 and 1951."

Fixed Wire-Wound Resistors for use in Telecommunication and Allied Electronic Equipment—Part 1: General Requirements and Tests

Pp. 17, B.S. 2111: Part 1: 1960. British Standards Institution, 2 Park Street, London, W.1. Price 10s.

MEETINGS

I.E.E.

8th May. "The Corona-Discharge and its Application to Voltage Stabilization", by E. Cohen and R. O. Jenkins, A.R.C.S., D.I.C., Ph.D.; "Impedance Frequency Characteristics of Glow-Discharge Reference Tubes", by F. A. Benson, D.Eng., Ph.D., and P. M. Chalmers, B.Eng.; "Comparison of Argon, Krypton and Xenon as Admixtures in Neon Glow-Discharge Reference Tubes", by F. A. Benson, D.Eng., Ph.D., and G. P. Burdett, B.Eng.

12th May. "The Place of Transistors in National Certificate Courses", discussion to be opened by B. F. Gray, B.Sc.(Eng.), and W. B. K. Ellis, B.Sc., at 6 o'clock.

15th May. "Circuits for the Banana Tube Colour Television Display System", by K. G. Freeman, B.Sc.; "Mechanical and Manufacturing Aspects of the Banana Tube Colour Television Display System", by H. Howden; "Appraisal of the Banana Tube Colour Television Display System", by K. G. Freeman, B.Sc., and B. R. Overton, B.Sc.; "The Banana Tube Display System—A New Approach to the Display of Colour Television Pictures", by P. Schagen, Ph.D.; "Colorimetry of the Banana Tube Colour Television Display System", by R. N. Jackson, and "The Banana Tube", by B. A. Eastwell, B.Sc., and P. Schagen, Ph.D.

17th May. "Air Traffic Control", by E. Eastwood, Ph.D., M.Sc., and B. J. O'Kane, Ph.D., B.Eng.

18th May. Annual General Meeting, followed at 6.30 by a lecture on "Experimental Investigation of Space", by J. A. Ratcliffe, C.B.E., M.A., F.R.S.

These meetings will be held at the Institution of Electrical Engineers, Savoy Place, Victoria Embankment, London, W.C.2, and will commence at 5.30, except where otherwise stated.

Brit. I.R.E.

17th May. "Television Wire Distribution Systems", discussion to commence at 6 o'clock at the London School of Hygiene and Tropical Medicine, Keppel Street, Gower Street, London, W.C.1.

The Society of Instrument Technology

11th May. "Control Mechanism in the Human Nervous System", by Dr. W. Grey Walter, M.A., Sc.D., to be held at 7 o'clock at Manson House, 26 Portland Place, London, W.1.

The Institute of Navigation

12th May. "Airborne Weather Radar", by Captain R. C. Alabaster, D.S.O., D.F.C., and Mr. P. L. Stride, to be held at 5.15 at The Royal Geographical Society, 1 Kensington Gore, London, S.W.7.

1961–1962 PRESIDENT OF E.R.A.

The British Electrical and Allied Industries Research Association has recently announced that Mr. James R. Beard, C.B.E., is to be its next President.

Mr. Beard is known internationally for his work in the electrical field and included in the many offices he has held in learned societies, institutions and other organizations is the Presidency of the Institution of Electrical Engineers, 1940–41.

STANDARD-FREQUENCY TRANSMISSIONS

(Communication from the National Physical Laboratory)

Deviation from nominal frequency for March 1961*

Date 1961 March	MSF 60 kc/s 1500 G.M.T. Parts in 10^9	Droitwich 200 kc/s 1030 G.M.T. Parts in 10^9
1	— 15·0	— 6
2	NM	— 6
3	— 14·9	— 6
4	— 14·9	— 5
5	— 14·9	— 4
6	— 15·0	— 4
7	— 15·2	— 4
8	— 14·8	— 4
9	— 15·0	— 2
10	— 15·1	— 2
11	— 15·4	— 2
12	— 15·5	— 3
13	— 15·2	— 3
14	— 14·8	— 3
15	— 15·0	— 3
16	— 15·1	— 3
17	— 15·1	— 3
18	— 14·8	— 2
19	— 14·9	— 1
20	— 14·7	— 1
21	— 14·3	0
22	— 14·2	+ 2
23	— 14·4	+ 1
24	— 14·1	+ 2
25	— 14·1	+ 3
26	— 14·8	+ 2
27	— 14·3	+ 3
28	— 14·2	+ 4
29	— 15·1	+ 4
30	— 15·4	+ 4
31	— 15·2	+ 4

* *Nominal frequency corresponds to a value of* 9 192 631 770 *c/s for the caesium* $F,m(4,0) - F,m(3,0)$ *transition at zero field. NM = Not measured.*

New Products

Sine and Square-Wave Generator

A sine and square-wave generator covering the frequency range 0·9 c/s to 120 kc/s with sine-wave distortion of 0·1% and a square-wave rise-time of less than 1 μsec has been announced by Waveforms Inc. U.S.A.

Output impedance is 600 Ω for both sine and square-wave signals and the output into a 600-Ω load is 10 V r.m.s. for sine waves or 10 V peak-to-peak for square waves. Both signals are controlled by a 5-position decade attenuator and a fine output control. A synchronization signal for oscilloscopes and counters is provided at terminals on the front panel.

Over the frequency range 9 c/s to 120 kc/s, dial accuracy is ±2% and frequency stability is 0·5%; at 1 kc/s, the step-attenuator accuracy is ± 2% and the fine-attenuator accuracy is ± 10%. The frequency response of the attenuator system is level ± 0·5 dB for all settings down to 100 μV.

Waveforms Inc.,
6 Cornelia Street,
New York 14, *N.Y., U.S.A.*

V.H.F. Attenuators

A range of attenuators which are intended for use at frequencies extending into the u.h.f. range and for incorporation into complete equipments have recently been announced by the components group of S.T.C.

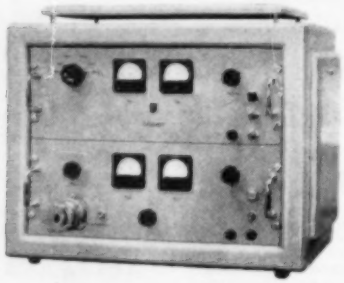
The range comprises 15 models with various maximum attenuations and attenuation steps, three different types of input and output connectors and a characteristic impedance of either 50 or 75 Ω. The attenuators range from 0–0·9 dB to 0–90 dB with attenuation steps from 0·1 dB to 10 dB, depending upon the models. The basic model number for all attenuators in the range is 74600.

Typical insertion-loss figures for the 0-dB setting at 50 Mc/s are 0·07 dB for the 0 to 0·9-dB attenuators (Models 74600-C, S and T), 0·12 dB for all other models.

Standard Telephones & Cables Ltd.,
Connaught House,
Aldwych, London, W.C.2.

High-Voltage Stabilized Supply

Designed for use with tubes requiring voltages in the range 1 to 2·25 kV, the Philips PP.4485 stabilized power supply is primarily intended for klystrons.



The following electrode supplies are available from this unit: (a) cavity, 1,000 to 2,250 V d.c. at 0 to 30 mA; (b) reflector, 0 to −550 V d.c. at 1 mA max.; (c) grid, 0 to −300 V d.c., and (d) heater, 6 to 7 V a.c.

Continuous indication of cavity voltage and current, together with the reflector and heater voltages, is provided by four built-in meters.

Provision is made for external modulation of reflector voltage and a safety device is incorporated to protect the klystron from any accidental change in operating conditions.

Distributed in the U.K. by
Research & Control Instruments,
207 *King's Cross Road, London, W.C.1.*

Silicon Power Transistors

The Semiconductor Department of A.E.I., enters the silicon power transistor field with the introduction of three Ediswan Mazda n-p-n diffused junction-type devices.

Known as types XC703, XC713 and XC723, they are intended for a wide variety of industrial applications in equipment operating at temperatures ranging from 175 °C to −65 °C. At 100 °C they offer a range of dissipation from 2–30 W and are available in Jedec cans.

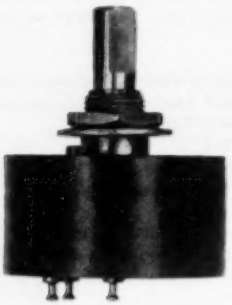
For the XC703, XC713 and XC723, the tentative characteristics are: maximum collector–base voltage 60 V, each; maximum collector current 1·5 A, 3 A and 6 A, respectively; static current amplification d.c. 15 (min.) and 75 (max.) for XC703 and 713 and 10 (min.) 50 (max.) for XC723; and collector–emitter saturation resistance d.c. 7 Ω (max.), 2·67 Ω (max.), and 2 Ω (max.), respectively.

Associated Electrical Industries Ltd.,
Semiconductor Department,
155 *Charing Cross Road, London, W.C.2.*

Plastic-Track Potentiometers

A range of potentiometers using homogenous conductive plastic resistance elements is now being imported into Great Britain from Ace Electronic Associates Inc., U.S.A.

Features claimed for the potentiometers include: a life expectancy of greater than 20-million cycles, low operational noise—equivalent to less than 25 Ω at 1 k Ω and 50 Ω at 10 kΩ, and a standard linearity of 0·5%.



The resistance values range, for the 1 $\frac{1}{16}$ in. diameter models, is 100 Ω to 250 kΩ and, for the $\frac{1}{2}$ in. diameter models, 100 Ω to 150 kΩ.

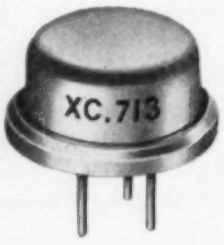
Power rating for the potentiometers is 2 W at 65° C, 2 kΩ and above and resolution is stated to be 0·005°.

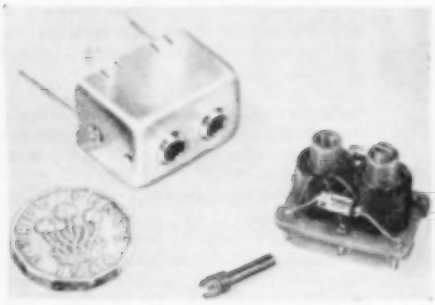
Distributed in the U.K. by
Scientific Furnishings Ltd.,
Poynton, Cheshire.

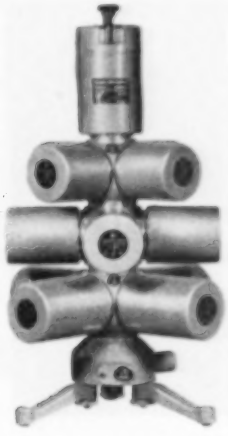
Miniature I.F. Transformers

A new range of miniature i.f. transformers designed primarily for transistorized receivers with intermediate frequencies within the range 450–480 kc/s have been introduced by Wireless Telephone Co.

The chief feature of the transformers is the side-by-side mounting of the coils which results in stable coupling and compact assembly. The ferrite cup cores are potted in epoxy resin to give good mechanical







rigidity with resistance to vibration. The can size is $\frac{11}{16}$ in. × $\frac{1}{2}$ in. × $\frac{11}{16}$ in. high.

Double-tuned transformers are used in the first two stages and a single-tuned unit in the third stage, and i.f. alignment is carried out by adjustable cores. In place of the normal threaded core a plain core is bonded to a polystyrene screw working in a threaded hole in the coil former.

The Wireless Telephone Co. Ltd.,
Hallamgate, Sheffield 8.

Portable Trigger-Pulse Generator

A compact portable trigger-pulse generator which has been designed to deliver fast high-voltage triggering pulses from a low source impedance has been announced by Winston Electronics.

Known as the pulse generator Type M.640, it provides in the free-running mode, positive or negative-going differentiated waveform pulses of amplitude 0–65 V

over the pulse recurrence frequency range 10 c/s to 100 kc/s. When triggered from an external signal source, the p.r.f. range is d.c. to 50 kc/s. The output impedance varies with the output-level potentiometer setting and is 500 Ω maximum. Rise time is less than 0·15 μsec.

A feature claimed for the generator is that the load capacitance has a negligible effect on the rise-time of the output signal; only slightly reduces the output-signal amplitude, and causes only a small shift in the p.r.f. The capacitive load required to reduce the output by 10% is 300 pF on the 1-μsec range and 0·1 μF on the 1-msec range.

The generator is powered from two internal 67·5 V batteries or from a separate mains power unit.

Winston Electronics Ltd.,
Shepperton, Middlesex.

Switch-Less Capacitance Standards

An unconventional switch-less capacitance-standards box has been introduced into the U.K. by Aveley Electric.

This is the Jahre system in which plug-in capacitance units are combined in 5- and 6-sided connector blocks which can be mounted in tiers. The capacitance range 10 pF to 10 μF is covered in 10-pF increments. The connector blocks are attached vertically to a base which houses the coaxial output socket and the standard capacitors are plugged in to the connectors horizontally. Unused sockets are covered by screening caps. Variable capacitors provide continuous coverage of 16 or 70 pF, with discrimination of 0·01 pF and 0·1 pF respectively.

Air dielectric is used in the sizes up to 400 pF which have loss factor below 1 in 10^5; series inductance is less than 0·06 μH per unit.

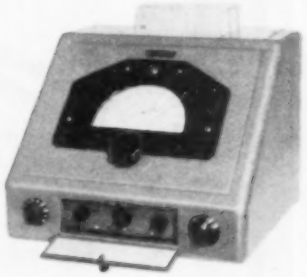
Each capacitor is adjusted to ±0·1% ±1 pF and the stability is guaranteed better than 0·05% per year. Temperature coefficients are +20 parts per million/deg. C for the units up to 400 pF and 30 parts/million for the larger units.

Distributed in the U.K. by
Aveley Electric Ltd.,
Ayron Road, Aveley Industrial Estate,
South Ockendon, Essex.

Moisture Meter

An instrument designed for the measurement of the moisture content of powdered, granulated and flake-like material has been introduced by Kappa Electronics.

Known as the model AB/55, it relies for operation on the change of permittivity of materials with moisture content and the superheterodyne principle. It comprises a variable-frequency oscillator (10 Mc/s mean), a fixed-frequency oscillator (10·15 Mc/s), a frequency changer or mixer stage, a detector (tuned to the 150-kc/s i.f.), a neon indicator, and a Perspex container for samples of the material being measured.



It is arranged for the Perspex container to sit between two plates of a capacitor in the variable-frequency oscillator tuning circuit. Initially, the scale of the variable capacitor, which is marked in arbitrary units, is calibrated at one or more points against samples with known moisture contents of the material being measured. Having calibrated the scale, the set-zero is adjusted for an i.f. of 150 c/s with the container empty, and in operation the change of the variable capacitor required to restore the i.f. to 150 kc/s is a measure of the permittivity of the material in the container and, therefore, of the moisture content.

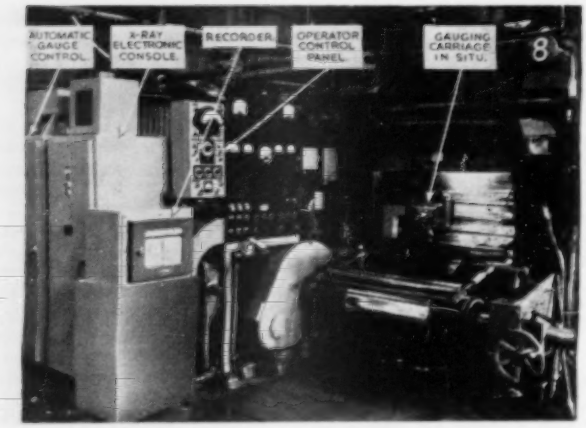
Accuracy is dependent upon a number of factors, especially the 'packing error' but, for normal materials of reasonably constant composition and texture, the accuracy can be expected to lie with 0·05–0·4% in the 0–20%, 0·2–0·5% in the 20–30%, 0·3–0·7% in the 30–40%, and 0·5–1% in above 40% moisture range.

Kappa Electronics,
159 Hammersmith Road, London, W.6.

Kerry-Omark Stud-Welding System

Kerry's (Ultrasonics) Ltd. and Omark Industries Inc., U.S.A. have announced the signing of a trade agreement for the exclusive sale of Omark stud-welding equipment by Kerry's in the U.K.

It is claimed that the Kerry-Omark system provides a unique extension to existing stud-welding techniques and gives a number of advantages over resistance and arc welding. Studs either threaded or unthreaded up to $\frac{1}{4}$ in. diameter can be welded to thin metals down to 0·02 in. thickness, without distortion or weld burn. No flux, ferrules or inert gases are necessary. Welds are produced by discharging a high



current at low voltage through a patented tip on the stud.

Welding machines are available for bench and portable use. The technique is suitable for welding aluminium, mild and stainless steel, brass, copper, titanium, etc.

Kerry's (Ultrasonics) Ltd.,
Tonbridge Road,
Harold Hill, Essex.

Miniature Integrating Gyro

Introduced recently by the Norden Division of United Aircraft Corporation, a miniature integrating gyro, Type RI-203, is now available in the U.K. through Ketay.

Features claimed for the gyro include a low drift rate—random drift is less than 0·02 deg./hr.; excellent day-to-day stability, 0·1 deg./hr. without trimming; and a long life of 3,000 hr. The low drift has been achieved with a spin motor turning at only 1,200 r.p.m.

The gyro measures 3·1 in. long by 2 in. diam., including heater and shields, and weighs 1 lb.

Ketay Ltd.,
Eddes House, Eastern Avenue West,
Romford, Essex.

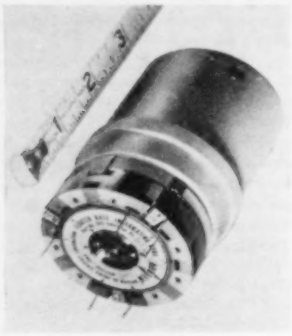
X-Ray Thickness Gauge

An X-ray gauge which is produced by the American Daystrom Company for the measurement of thickness variation of rolled and extruded metals is now available in the U.K.

The gauge is known as the XactRAY and consists of an X-ray generator which directs controlled beams of X-rays through the materials being measured, the X-ray incident unit or pick-up and an electronic control and display unit which converts the output from the incident unit into meter readings. Dimensional changes are displayed as a deviation from the pre-set nominal thickness.

Five models of the gauge give a range from 0·00035 in. for aluminium foil to 2-in. steel plate; any one model has a thickness range of up to 200 : 1 (a 60-kV X-ray generator is suitable for 0·01–2 in. of aluminium). Accuracy is better than ±1% of the thickness being measured.

Daystrom Ltd.,
Gloucester.



Infra-red Photocell

Mullard have developed the photoconductive cell type ORP13 which it is claimed has a higher detectivity than any other photocell with infra-red applications yet commercially available in the world. It is intended for applications involving infra-red spectroscopy and research into the properties of materials.

The cell, which has a sensitive area of approximately 3 sq. mm of indium antimonide, is attached to its own Dewar flask and is cooled by liquid nitrogen to a temperature of 77 °K (−196 °C). This cooling system enables very good noise figures to be obtained and its efficiency is such that the cell can be used for 45 min. without refilling the flask.

The spectral range of the ORP13 is from visible to about 5·4 μ, the peak occurring at between 4·5 and 5 μ. Its sensitivity to monochromatic radiation is 15 mV per μW of incident radiation.

The noise equivalent power (n.e.p.) per unit bandwidth (1 c/s), which is a measure of the smallest radiation power conveniently detectable by the cell, is better than $1{\cdot}3 \times 10^{-11}$ W. This figure gives an area-normalized detectivity D* of better than $1{\cdot}3 \times 10^{10}$ cm W^{-1}.

Mullard Ltd.,
Torrington Place, London, W.C.1.

* D is a figure of merit for photocells used for detecting low-level radiations and represents a basis for the comparison of cells of different sensitive areas and different materials. Unlike n.e.p., D is independent of sensitive area since it is the square root of the sensitive area divided by n.e.p.

Miniature Condenser Microphones

Bruel & Kjaer have recently released two ½-in. condenser microphones which have been designed primarily for sound measurements. Two cathode followers have also been developed for the microphones.

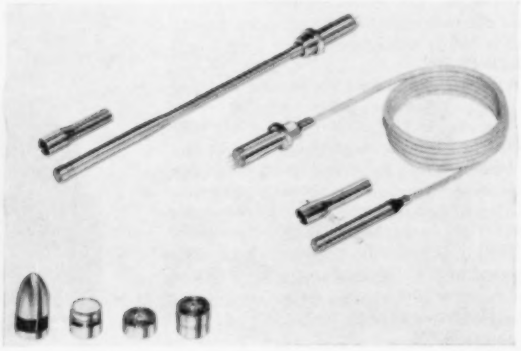
Microphone type 4133 has a linear frequency range of 20 to 40,000 c/s within approximately ±1 dB for sound waves impinging perpendicularly on the diaphragm and is intended for use in free-field measurements.

Microphone type 4134 has a linear frequency response from 20 to 20,000 c/s (within ±1 dB) and is intended for use in a diffused sound field or under acoustical pressure conditions.

Both microphones are capable of measuring sound pressure levels above 160 dB, reference 2×10^{-4} μbar, and can be used for accurate measurements well above 150 °C.

Among the many ancillaries for these microphones is a new windscreen called the nose cone UA0052. When this nose cone is mounted in place of the normal protecting grille, the type 4133 becomes an omni-directional microphone with a flat frequency response for all angles of incidence to above 14 kc/s.

B & K Laboratories Ltd.,
4 Tilney Street, London, W.1.



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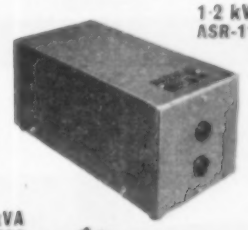

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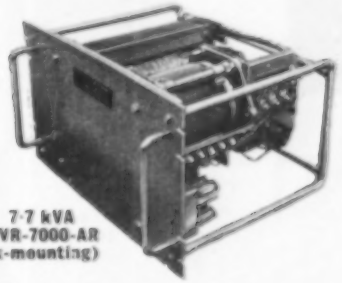

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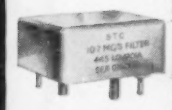

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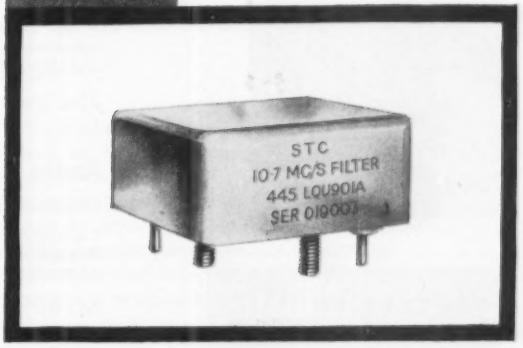

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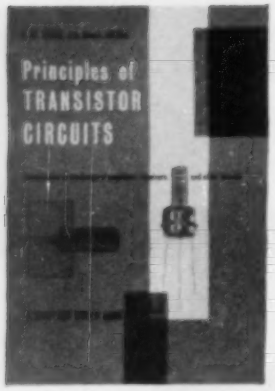

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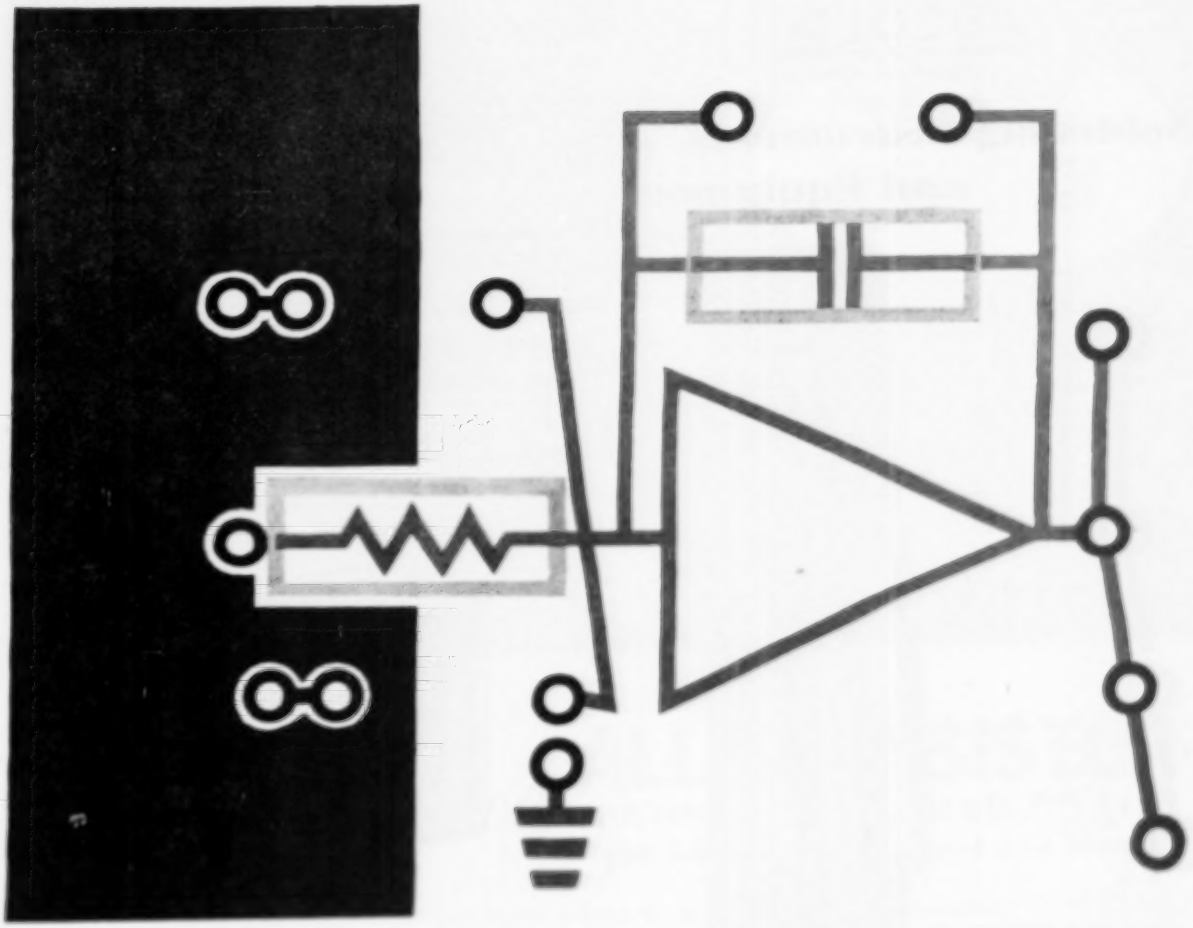

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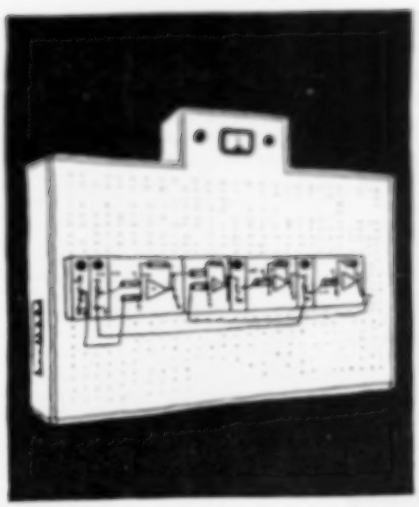
for demonstrating computing techniques and for research



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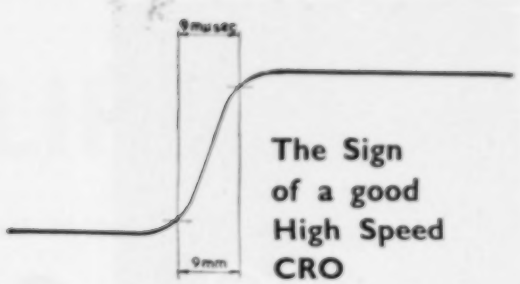

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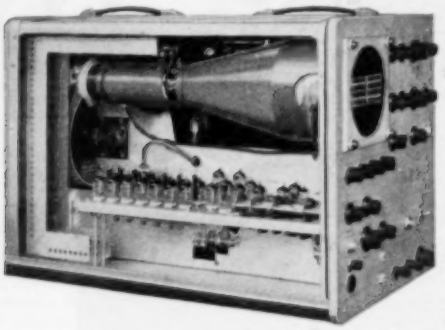

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Rear view of a Multiplexer-Demultiplexer Shelf, Type 250 Model 1, showing 2 Transistorized Power Suppliers and Changeover Relay.

NORTHERN RADIO

New ALL-TRANSISTOR

MULTIPLEXER and DEMULTIPLEXER

Type 248 Model 1 Type 249 Model 1

The new Multiplexer, Type 248 Model 1 (functional replacement for Multiplexer TD97-FTG-2), and Demultiplexer, Type 249 Model 1 (functional replacement for Demultiplexer TD98-FGR-3) are intended for use with twin-channel, single-sideband radio circuits operating in the high-frequency range. Their purpose is to derive two voice-frequency circuits from each of the radio channels. By means of frequency division multiplexing, the radio bandwidth from 200 to 6000 cps is divided into two transmission circuits, each with a bandwidth from 375 to 3025 cps. Four such vf circuits are derived from the twin-channel radio, and these are used to transmit carrier telegraph signals or to provide telephone or facsimile service.

The Multiplexer and Demultiplexer are designed to slide into the Northern Radio Type 250 Model 1 Shelf, which accommodates two each Multiplexers or Demultiplexers, or one each Multiplexer and Demultiplexer.

Two Multiplexers, Type 248 Model 1, are required for full utilization of the capacity of a radio transmitter. One is used to transmit telegraph, telephone, or facsimile signals from two vf circuits to the radio channel designated as sideband A. The second Multiplexer performs the same function for sideband B. In this way four vf circuits are applied to the twin-channel radio transmitter.

Two Demultiplexers, Type 249 Model 1, are required for full utilization of the capacity of a radio receiver. One is used to receive telegraph, telephone, or facsimile signals for two vf circuits from the radio channel designated as sideband A. The second Demultiplexer performs the same function for sideband B. In this way four vf circuits are derived from the twin-channel radio receiver.

The Multiplexer and Demultiplexer are transistorized equipments, including necessary bandpass filters, line amplifiers, carrier frequency sources, modulators and attenuators. The Multiplexer requires a nominal 14 volts DC at 125 milliamperes; the Demultiplexer, approximately 200 milliamperes at the same voltage. The power supply is normally provided from the Northern Radio Power Supply, Type 223 Model 1, which is plugged into the rear of the Type 250 Model 1 Shelf.

WRITE FOR FURTHER INFORMATION ON YOUR LETTERHEAD TO DEPARTMENT ET-5

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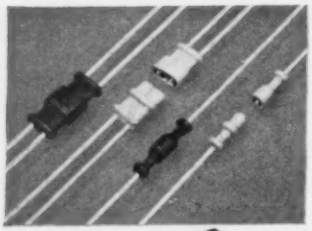
Mouldings and Electronic Components

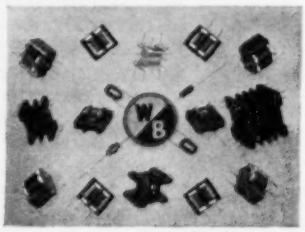


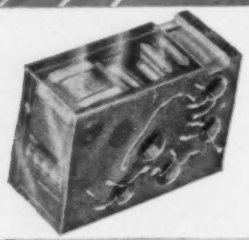
PLUGS AND SOCKETS

The plugs and sockets shown are for use in underwater junctions or outside installations. Moulded in Alkathene the loading capacity is 350 volts at 5 amps. max.

MINIATURE BOBBINS

Illustrated are a range of moulded bobbin assemblies for small transformers and chokes. Moulded in Polystyrene or Polythene, they are available in various colours.



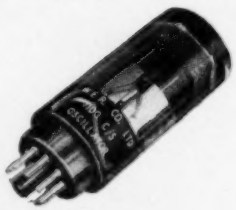


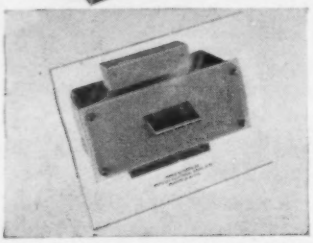
P.O. TRANSISTOR AMPLIFIER

This P.O. Transistor Amplifier moulding is another example of potted components, the illustration is of a development of a miniature version of the amplifier 100A, this model utilising transistors to replace the miniature valves.

PLUG-IN OSCILLATOR

This fixed frequency oscillator is constructed on a standard octal base and encapsulated in epoxy resin. Output: 10 mw into a 600 ohm load. Frequency: as required within the range 700-2000 c.p.s. sinusoidal.





SCREEN LINE TRANSFORMER

This is an astatic wound A.F. screened line transformer with the windings encapsulated in epoxy resin. The insulation of the "line" winding provides isolation against voltages of 30 kV RMS.

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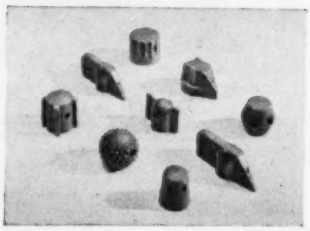

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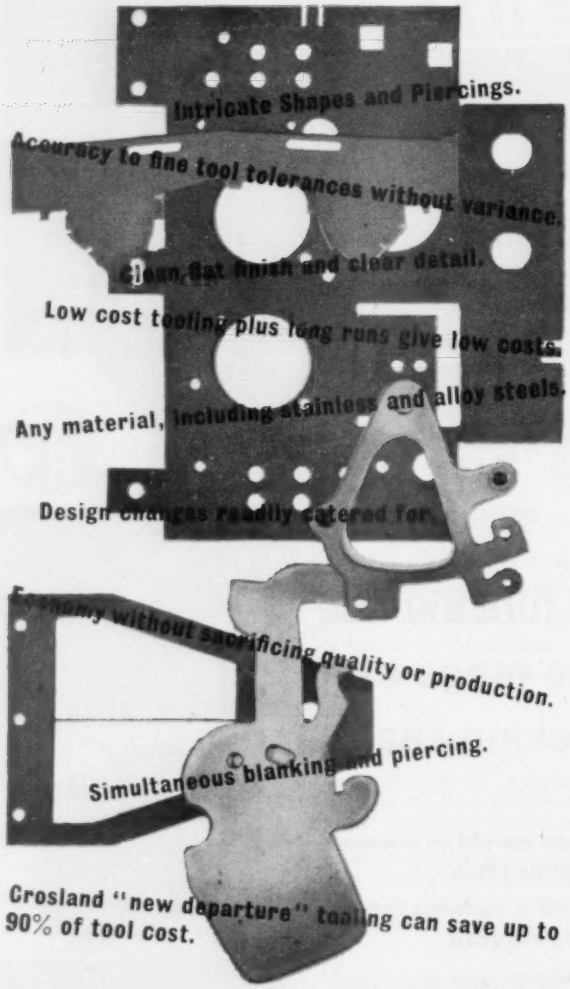
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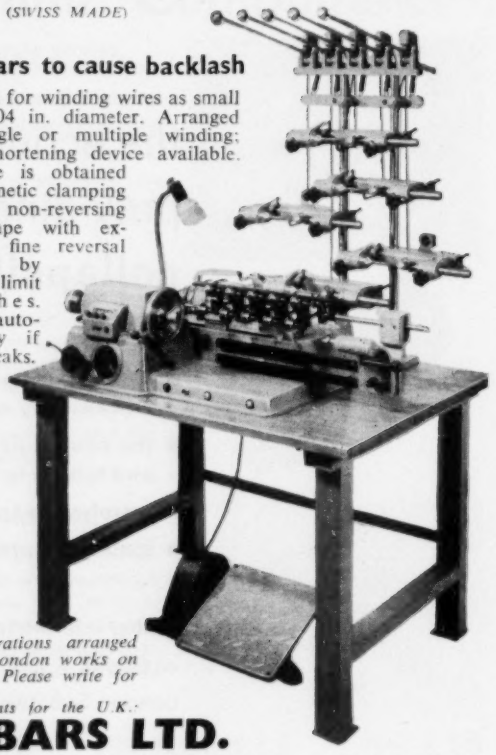

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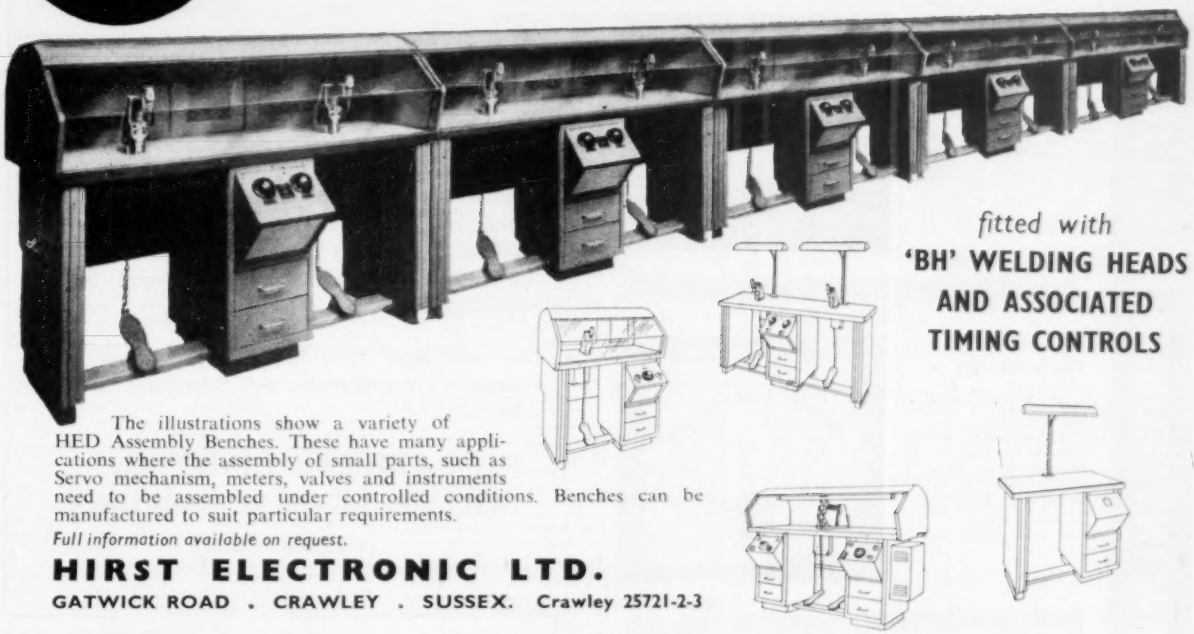

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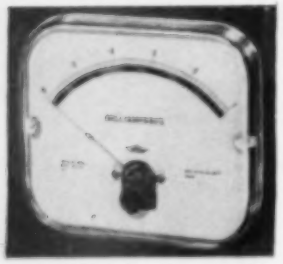

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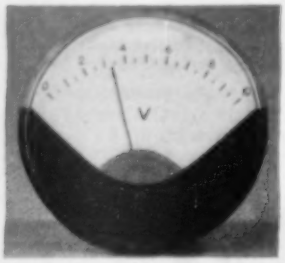

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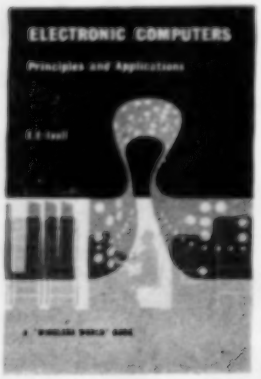

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Coventry

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W. L. CHINN,
Director of Education.
[1461

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APPLY stating age, qualifications and experience and giving the names of three persons to whom reference may be made to the Secretary, 80 Rodney Street, Liverpool, 1, as soon as possible.
[1460

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Personnel Officer,
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Applications may be made in the strictest confidence to the Personnel Manager, Box No. 5004, c/o "ELECTRONIC TECHNOLOGY", stating qualifications, experience, age, and salary required, etc.

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The A.I.D. approved type 362 flux, incorporated in Ersin Multicore 5-Core Solder, is very effective on heavily oxidised surfaces and often allows the use of a lower tin content alloy. Ersin Multicore 5-Core Solder is supplied on 7 lb. reels in 9 standard gauges and 6 alloys. Even gauges from 24-34 s.w.g. are available in 2 alloys on 1 lb. and ½ lb. reels.

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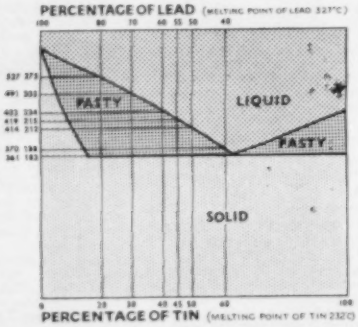
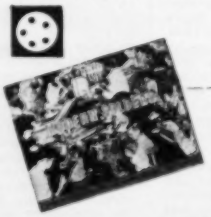
The diagram shows that all the standard alloys of Ersin Multicore Solder have a plastic range, i.e., on heating they are pasty between the solid and liquid states. Practical experience has shown that there are advantages in having a plastic range, e.g. for tag jointing where the use of 60/40 alloy obviates fractures which may occur with other alloys where there is slight vibration while the solder is setting solid.



(left) after 10,000 joints with Savbit. (centre & right) after 1,000 and 7,500 joints respectively with standard tin/lead alloy.

PRINTED CIRCUITS

Leaflet P.C.L. 101 gives full details of a complete soldering process developed by Multicore Laboratories for printed circuits.



SAVBIT TYPE 1 ALLOY

Made under sole British Licence of Patent No. 721,881.

Savbit Type 1 alloy was developed after extensive research in the Multicore Laboratories into the main causes of bit wear. It incorporates a small percentage of copper which prevents absorption of copper from the bit into the solder alloy. After prolonged tests, it was found that the life of solder bits was increased by up to 10 times. The speed of soldering is not affected.

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PUBLICATIONS

Laboratory engineers and technicians are invited to write on their company's letter-heading for the latest edition of Modern Solders. It contains data on melting points, gauges, alloys, etc.

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Index to Advertisers

	PAGE
Acbars Ltd.	22
Adcola Products Ltd.	18
Appointments Vacant	24, 25
Associated Electrical Industries Ltd.	10, Cover iii
Automat Ltd.	20
Avo Ltd.	3
Box No. 5004	25
British Insulated Callender's Cables Ltd.	Cover ii
Brush Crystal Co. Ltd.	8
Collico Ltd.	21
Crosland, Wm., Ltd.	22
Ferranti Ltd.	6
Gabriel Manufacturing Co. Ltd.	14
Gardners Radio Ltd.	1
Hirst Electronic Ltd.	23
Hughes International (U.K.) Ltd.	5
Iliffe Books Ltd.	16, 24
International Computers & Tabulators Ltd.	2
Interplas	7
Labgear Ltd.	24
Leland Instruments Ltd.	Cover iv
Lyons, Claude, Ltd.	13
Marconi's Wireless Telegraph Co. Ltd.	9
Measuring Instruments (Pullin) Ltd.	24
Metropolitan Overseas Supply Corp.	26
Micanite & Insulators Co. Ltd., The	23
Mullard Ltd.	12
Multicore Solders Ltd.	26
Murex Ltd.	25
Nagard Ltd.	18
Northern Radio Co. Inc.	19
Oxley Developments Co. Ltd.	22
Parsonage, W. F., & Co. Ltd.	26
Short Brothers & Harland Ltd.	17
Standard Telephones & Cables Ltd.	15
Steatite & Porcelain Products Ltd.	4
Whiteley Electrical Radio Co. Ltd.	20

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NEW VERY HIGH SLOPE SCREENED HF PENTODE

EDISWAN MAZDA 6F24

The 6F24 has a slope of 15 mA/V at 10 mA anode current and a high "Figure of Merit," both results achieved by incorporating the new frame grid technique. This new screened HF Pentode enables improved performance to be obtained as an IF amplifier in a.c. or a.c./d.c. television receivers.

Heater Voltage (volts)	V_h	6.3
Heater Current (amps)	I_h	0.3

TENTATIVE RATINGS AND DATA

Maximum Design Centre Ratings

Anode Dissipation (watts)	$p_{a\ (max)}$	2.5*
Screen Dissipation (watts)	$p_{g2\ (max)}$	0.8*
Anode Voltage (volts)	$V_{a\ (max)}$	250
Screen Voltage (volts)	$V_{g2\ (max)}$	250
Heater to Cathode Voltage (volts rms)	$V_{h\text{-}k\ (max)\ rms}$	150†
Control Grid to Cathode Resistance (megohms)	$R_{g1\text{-}k\ (max)}$	0.6‡

*With grid to cathode resistance not exceeding 10 kΩ.
†From cathode to higher potential heater pin.
‡With $p_{a\ (max)} = 2$ W; $p_{g2\ (max)} = 0.5$ W; and assuming a common anode and screen decoupling resistance, ≮ 2.2 kΩ ± 10%.

Inter-Electrode Capacitances (pF)§

Input Capacitance	c_{in}	8.8
Output Capacitance	c_{out}	2.6
Grid 1 to Anode	$c_{g1\text{-}a}$	0.006
Grid 1 to Grid 3	$c_{g1\text{-}g3}$	0.1
Grid 1 to Grid 2	$c_{g1\text{-}g2}$	2.0
Grid 1 to Cathode	$c_{g1\text{-}k}$	6.2
Grid 2 to Anode	$c_{g2\text{-}a}$	0.15
Grid 3 to Anode	$c_{g3\text{-}a}$	0.47

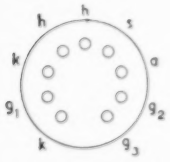
§Measured in fully shielded socket, without can.

TYPICAL OPERATION

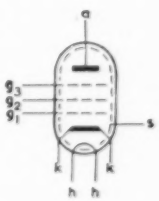
Anode Voltage (volts)	V_a	170
Screen Voltage (volts)	V_{g2}	170
Self Bias Resistance (ohms)	R_k	150
Anode Current (mA)	I_a	10
Screen Current (mA)	I_{g2}	2.7
Mutual Conductance (mA/V)	g_m	15
Inner Amplification Factor (g_1 to g_2)	$\mu_{g1\text{-}g2}$	65
Equivalent Grid Noise Resistance (ohms)	R_{eq}	370
Input Loss at 38 Mc/s (Pins 1 and 3 strapped) (kΩ)	$r_{g1\text{-}k\ (w)}$	8.5
Working Input Capacity** measured at 38 Mc/s (pF)	$c_{in\ (w)}$	13.7
Change in Input Capacity produced by biasing valve to cut-off, measured at 38 Mc/s (pF)	$\Delta c_{in\ (w)}$	3.4
Figure of Merit†† (Valve only) (Mc/s)		375
Effective Figure of Merit (Valve and Circuit) (Mc/s)		220

**Inter-electrode capacity with holder capacity balanced out.
††Given by $\frac{g_m \times 10^3}{2\pi\sqrt{c_{in\ (w)}\, c_{out}}}$ see "Aspects of Design" No. 1 for further details (Wireless World July 1958).

Base: B9A (Noval) **Mounting Position:** Unrestricted



VIEW OF FREE END



Maximum Dimensions (mm)

Overall Length	56
Seated Height	49
Diameter	22·2

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Radio and Electronic Components Division
Technical Service Department
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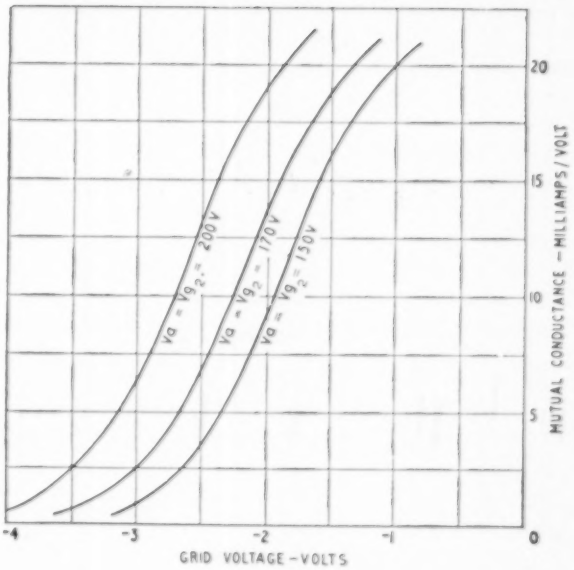
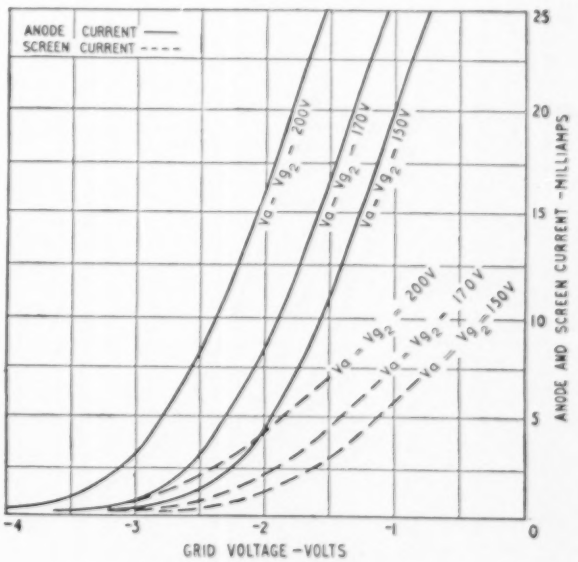
Tentative Characteristic Curves of Ediswan Mazda Valve Type 6F24





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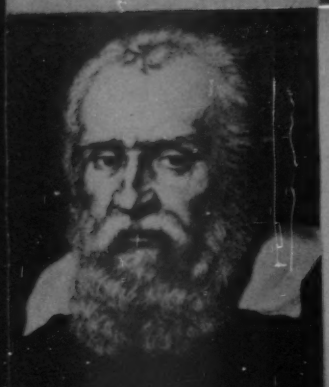
48 inch telescope at Asiago. University of Padua



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